Universidade de Aveiro Departamento de Electrónica, Telecomunicações e Informática
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dissertação apresentada à Universidade de Aveiro para cumprimento dos requisitos necessários à obtenção do grau de Mestre em Engenharia Electrónica e Telecomunicações, realizada sob a orientação científica do Dr. Dinis Magalhães dos Santos, Professor Catedrático do Departamento de Electrónica, Telecomunicações e Informática da Universidade de Aveiro

Dedico este trabalho aos meus pais, irmã e em especial à Ana

o júri	Professor Doutor Dinis Gomes de Magalhães dos Santos
	Doutor Domingos da Silva Barbosa
presidente	Professor Doutor José Carlos da Silva Neves

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palavras-chave

Rádio Astronomia, Electrónica de RF, Microondas, Simulação, Desenho, Implementação de Circuitos de RF.

resumo

Este trabalho contextualiza-se no âmbito de uma experiência de Mapeamento da Emissão Galáctica, para tal está a ser desenvolvido um sistema capaz de recolher dados galácticos a 5 GHz com o objectivo de caracterizar a Radiação Cósmica de Fundo (FRCM). Para o sistema de recolha de dados do Hemisfério Norte está a ser desenvolvido um receptor para o efeito. Trata-se de um polarímetro heterodino a 5 GHz com elevado ganho a Frequência Intermédia (FI) que utiliza a última tecnologia de RF a funcionar a 600MHz com uma largura de banda de 200 MHz que alimenta um correlador totalmente digital de quatro canais. Anterior a IF encontra-se um sistema de conversão de RF (5 GHz) para FI e um filtro de rejeição de imagem a esta frequência. O primeiro componente do cadeia do receptor, logo a seguir ao OMT (Orthomode Transducer) é um amplificador de muito baixo ruído (LNA). Este trabalho descreve o pré amplificador de FI com filtro passa-banda, um amplificador de FI com controlo digital de atenuação, um conversor para banda base com modulação em fase e quadratura, um filtro passivo de microondas a 5 GHz, uma pequena introdução do desenho previsto do LNA e uma abordagem ao hardware desenvolvido para o correlador digital. São apresentadas as opções de desenho e dificuldades encontradas no desempenho do circuito, juntamente com os resultados de simulação e experimentais obtidos para um protótipo.

keywords RF electronics, Microwaves, Simulation Design and Circuit Implementation

abstract

In the context of the Galactic Emission Mapping collaboration, a galactic survey at 5GHz is in preparation to characterize the galactic foreground to the Cosmic Microwave Background Radiation. For the North sky survey, a new receiver is being developed. This is a 5GHz heterodyne polarimeter with a high gain IF chain using the latest RF technology working at 600MHz central frequency that feeds a four channel digital correlator. Prior to this chain is the first down conversion from RF (5 GHz) to IF (600 MHz) and a microwave passive filter also design and implemented, and a very Low Noise Amplifier (LNA). This thesis describes the preamplifier/band-pass filter, the digitally controlled amplifier, the frequency converter to zero-IF, a microwave passive filter, a introduction on the LNA design and a briefly description of the hardware of the digital correlator. Design options and constraints are presented along with the simulations and experimental results of a circuit prototype.

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1. Introduction

1.1. Motivation

History

"The most profound and the most fruitful that physics has experienced since the time of Newton", this was the way how Einstein described the work of James Maxwell. James Maxwell developed the electric and magnetic forces theory described in his famous equations. These equations already announced the existence of radiation, later known as electromagnetic radiation. With this new achievements occurred to Heinrich Hertz to demonstrate the existence of electromagnetic radiation building an apparatus that could transmit and receive electromagnetic waves of about 5 meters in length. Once Hertz had demonstrated the existence of electromagnetic radiation, the possibility of receiving such radiation from celestial objects may have occurred to many scientists. Edison seems to be the first on record to have proposed an experiment to detect radio waves from the Sun. The evidence of this is a letter sent in 1890 to Lick Observatory by Kennelly, who worked in Edison's laboratory. The detection of radiation from Sun was challenging to several Physicists, but unfortunately all the attempts failed. The principal cause was ionosphere discovered by Heavyside in the twenties that demonstrated its existence and absorption of low frequency radiation (20 MHz). Ionosphere was in fact a hard obstacle to overcome, the incident waves in this layer are reflected, coming either by the outer space or from earth. But this difficulty revealed very useful for long distance communications. By reflection in ionosphere longer distances were achieved and was possible to communicate farther away. Marconi was the first to develop a capable system to emit and receive signals beyond an ocean. This transatlantic transmission was the culmination of several trials of "hertzian" communication that started with Hertz simple experiment. But one had to wait for the appearance of the first global radio communications to see the rise of a new activity — Radio Astronomy (RA). Nowadays, one of the most important branches in astrophysics, the establishment of RA was the result of a sequence of accidental discoveries made largely by radio engineers and amateurs pushing the envelope of advances in shortwave IT technology.

In 1931-1935 the works of radio engineer Karl Jansky, charged by Bell Telephone Laboratories to investigate using "short waves" for transatlantic radio telephone service. Jansky was assigned the job of investigating the sources of static that might interfere with radio voice transmissions. He eventually figured out that the interfering radiation was coming from the Milky Way. Jansky wanted to follow up on this discovery and investigate the radio waves from the Milky Way Galaxy in more detail. These were the first steps in RA that revealed the need of having larger dish antennas. Due to the Great Depression Jansky did not have the chance to continue with his investigations. Grote Reber, an amateur radio, was the follower of Jansky, after reading papers from Jansky he built a telescope (9,5 m dish antenna made in wood and iron tuned for UHF – 160 MHz) in the backside of his house. With this new development he urged once again RA, announcing the first Milky Way maps.

In 1954, E. Purcell discovered the Hidrogen Line in the Universe at 1420 MHz. These frequency values are very important for radio astronomers and still are protected by law. All these innovations in physics and the rise of RA result in great advance in Astrophysics, more properly, the discovery of Cosmic Microwave Backgroung Radiation (CMBR).

CMBR

The beginning of the Universe, its evolution and geometry represent one of the great challenges of Humanity. The Big-Bang features the first Universe steps. At the beginning there was light (photons!). The primordial Universe was hot, matter was completely ionized and its dynamics was governed by a huge radiation bath. Actually the Universe is expanding, allowing to determine that in the past it was smaller. As the Universe expanded and cooled, the atoms formed (380 000 years after the Big-Bang) and this radiation bath could finally escape carrying the imprints of the forming Large Scale Structure (the small temperature fluctuations are like a xerox copy of the matter fluctuations at the time). Nowadays, with almost 14 thousand million years of Cosmic History, this radiation bath forms the Cosmic Microwave Background Radiation (CMBR). The temperature of CMBR is about 2,7 K (Kelvin) and it is responsible for 1% of the noise in our domestic TV receivers. This radiation is every where, surrounding us, but with very low amplitude to be detected with normal devices, thus in order to see this radiation one needs high sensible devices to detect it. The CMBR presents small fluctuations o about ~50~80 μK these variations involved are in temperature and in polarization, and represent the matter irregularities that have grown up with time and became the Galaxies it is possible to see today. Other characteristic of CMBR is the spectrum, which is described by a blackbody spectrum. In figure 1 is the spectrum of a blackbody demonstrated by Max Planck.

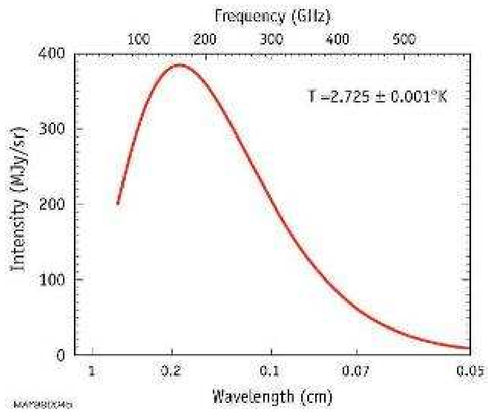

Figure 1 – Blackbody Spectrum Description in "Physics World December"

This background radiation had in fact been predicted years earlier by George Gamow as a relic of the evolution of the early Universe. This background of microwaves is the cooled fraction of the early fireball - an echo of the Big Bang.

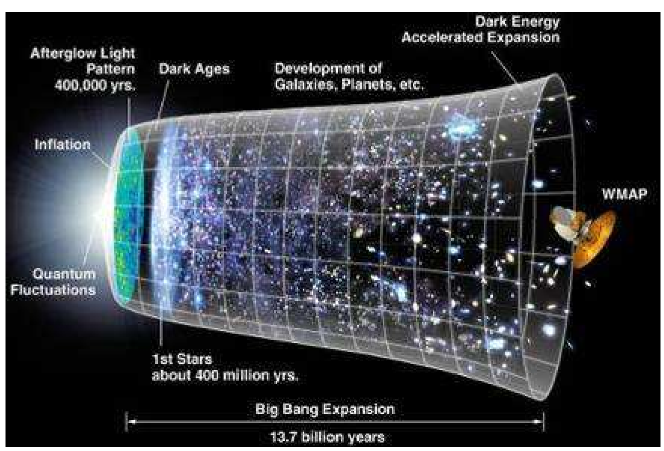

Figure 2 – Time Line of the Universe

Gamow did some calculations of the conditions of primordial Universe, although they were not right it served as the first attempt to understand the galaxy formation. His calculations determined that the origin of the Universe was extremely hot. Later he and his collaborators calculated that the density of the primordial Universe radiation was greater then the matter density [4]. In 1949 Alpher and Hernan studied the radiation temperature evolution from the primordial Universe till the time and announced a 5 K value for temperature radiation. In 1953, Alpher Follin e Hernan studied deeply the Universe temperature radiation but with no results. In 1964 Doroshkevich and Novikov determined the radiation relic and verified that it had a spectrum equal to the one described by a blackbody spectrum. In 1965, two young radio astronomers, Arno Penzias and Robert Wilson, almost accidentally discovered the CMB using a small, well-calibrated horn antenna. It was soon determined that the radiation was diffuse, emanated uniformly from all directions in the sky, and had a temperature of approximately 2.7 Kelvin (i.e. 2.7 degrees above absolute zero). Initially, they could find no satisfactory explanation for their observations, and considered the possibility that their signal may have been due to some undetermined systematic noise. Penzias went on to work at Bell Labs in Holmdel, New Jersey where, with Robert Woodrow Wilson, worked on ultra-sensitive cryogenic microwave receivers intended for radio astronomy observations. In 1964, on building their most sensitive antenna/receiver system, the pair encountered radio noise which they could not explain. It was far less energetic than the radiation given off by the Milky Way, and it was isotropic, so they assumed their instrument was subject to interference by terrestrial sources. An examination of the microwave horn antenna showed it was full of pigeon droppings. It soon came to their attention through Robert Dicke and Jim Peebles of Princeton that this background radiation had in fact been predicted years earlier by George Gamow.

Recently CMBR research is made by COBE – Cosmic Background explorer satellite, developed in Goddard Space Fligth Center from NASA. COBE measures primordial Universe microwave radiation [6]. Recently was awarded the Physics Nobel Prize to George Smoot due to his investigations that revealed results that allowed him to create the first CMBR map obtained from the Differential Microwave Radiometer (DMR) on board of COBE. Other project gave rise to more detailed picture of infant Universe, it was the WMAP satellite launched in 2001. This new information permits identification of when were formed the first stars and provides clues about what happened in the 10-13 second of Universe. Actually CMBR investigation continues with ESA Planck Surveyor Satellite (launch previewed for 2007) and by ESO/NRAO 64 12 m antennas Atacama Large Milimeter Array (ALMA). Both surveys are in an experiencing phase, it will work for several frequencies from 30 GHz to 900 GHz.

CMBR was detected for the first time in 1965 and mapped for the first time with the DMR instrument aboard COBE satellite in 1992. George Smoot, PI of COBE/DMR, was awarded the Physics Nobel Prize in 2006 for this discovery.

1.2. Overview

GEM

All the work described in this thesis fits Radio Astronomy Experiences and is a part of the development of a radio telescope. All this integrates the research project designated GEM – P (Galactic Emission Mapping – Portugal). The used radio telescope is set by antenna, a receiver, acquisition data system and a data storage system and antenna control. GEM-P has a direct relation with CMB. In figure 3 are presented the maps obtained from the COBE and WMAP satellites.

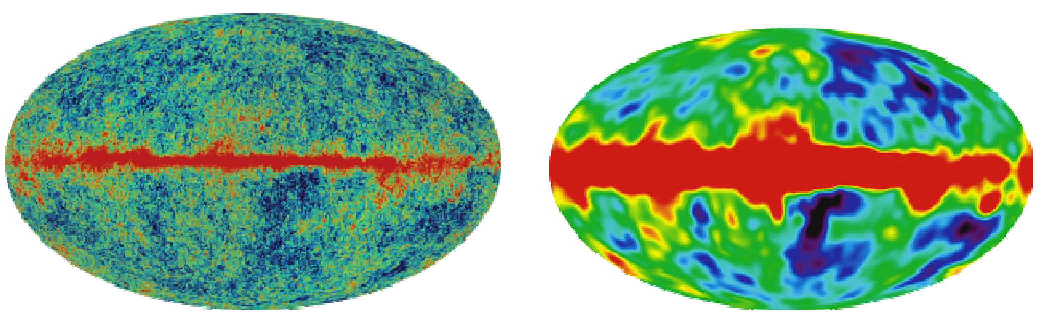

Figure 3 – CMBR maps obtained by WMAP (left) and by COBE (right)

It is possible to note that both maps suffer from an equal problem that is the presence of the line in the middle. These maps show the signal of CMBR plus the signal from our galaxy, denoted by the red line. As was said before, fluctuations of temperature together with polarizations changes determine the best proof of the beginning of the Universe. These results only contain information of temperature. In order to get clean and complete maps of CMBR, like Planck, there is a need to know also information about polarization. Devices like this are need to have high sensitivity e calibration to gather the polarized signal. Once this is done it is possible to subtract the radiation from our galaxy and obtain a clean map of CMBR, lifting the veil to the CMBR, like in figure 4.

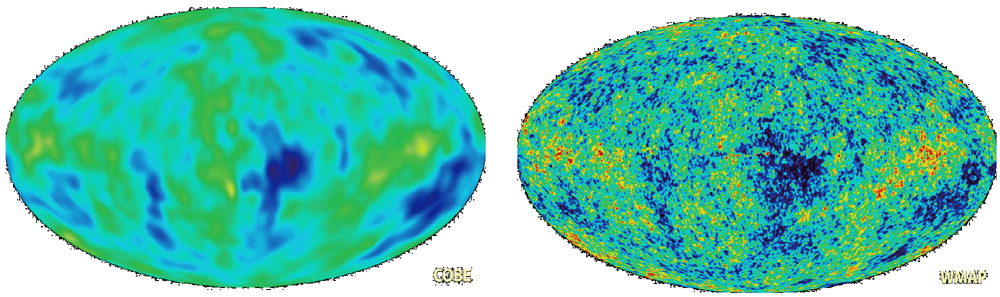

Figure 4 – CMBR maps without Milky Way radiation

This challenge took Prof. George Smoot (Group of Astrophysics in LNBL – Lawrence Berkeley National Laboratory, USA) and his collaborator Sérgio Torres. The main objective of GEM is to quantify the galactic contamination. GEM will map, with high sensitivity and absolute calibration the sky (and the Milky Way). Several scientific institutions are connected to this project, namely: Instituto de Telecomunicações – Pólo de Aveiro, Portugal; CENTRA – Centro Multidisciplinar de Astrofísica, Portugal; LNBL, INPE – Instituto Nacional de Pesquisas Espaciais (which is the NASA equivalent in Brazil) and Universitá di Milano. Italy. GEM is divided in two groups one in the South Hemisphere (Brazil) and other in North Hemisphere (Portugal – GEM-P). Together will survey approximately 85% of the sky. In figure 5 is a preview of the desired results.

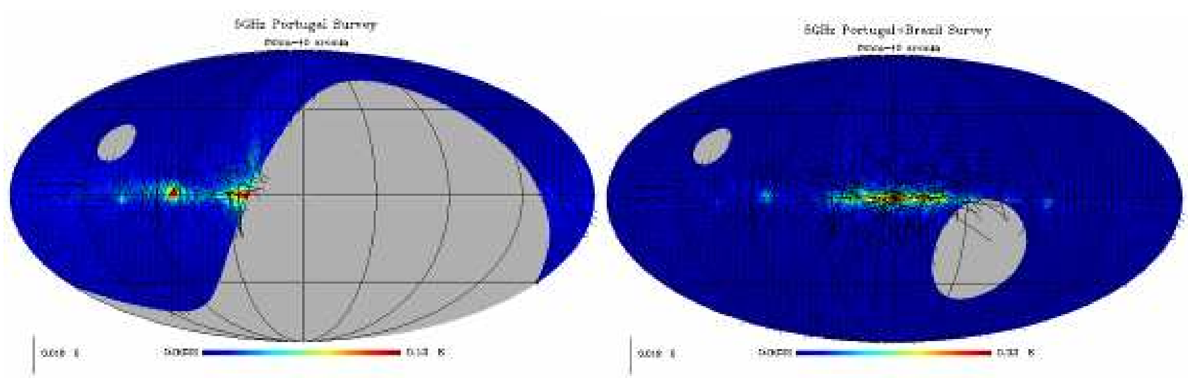

Figure 5 – GEM sky coverage expected

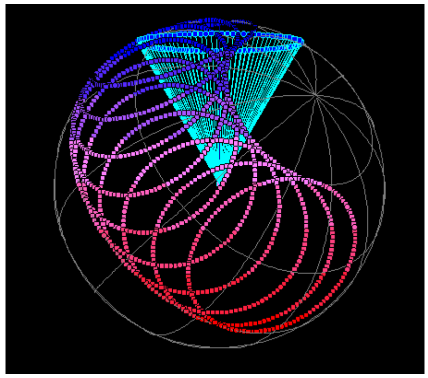

Figure 6 – Strategy of scan

Basically GEM-P will map the sky at 5 GHz, the strategy of scan will be an antenna pointed 30º from zenith rotating at 1 rpm. This strategy will avoid 1/f noise and electronic noise generated by the receiver. Together with this is the Earth rotation that will avoid noise generated by the atmosphere. The antenna used is a Cassegrain Vertex RSI high performance nine meter dish antenna placed in a low RFI (Fonseca et al.2006) site (long.7º52' Lat. 40º11'). The receiver will be located right next to the feed, as near as possible to it. It is a very low noise receiver works with very low signal levels (sub miliKelvin).

Frequency

At 5 GHz the largest contribution of contamination from our Galaxy is due to synchrotron emission as it is verified in figure 7.

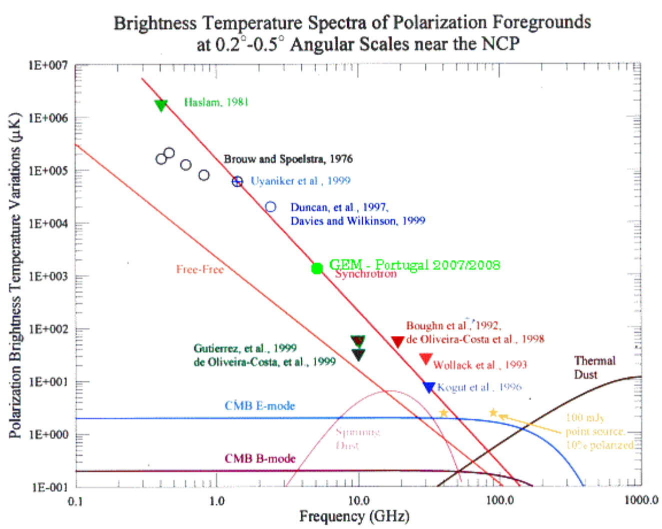

Figure 7 – Temperature values spectrum

In figure 7 the GEM-P is placed at 5 GHz which corresponds to a Brightness Temperature of 1 mK for Synchrotron, this value indicates that the receiver needs to detect signals of this order and also keep information of little signal variations. Resuming it will have a resolution lower than 1mK.

1.3. Thesis Organization

Throughout this thesis will be describe the area all the RF electronics and microwaves that involve the development of the receiver to use in this project. The receiver is a radiometer / polarimeter, the analysis of the collected data will be made in a digital level being the first of this type – full digital correlator - (it is a functional block where the correlation and integration of the signals gathered from the antenna are manipulated in the digital domain) and with a conversion to base band using modulation in phase (I) and Quadrature (Q). More properly it will be described in this thesis the development of a image rejection filter, an pre amplifier at IF with filtering, a converter to base band with I and Q modulation, the respective local oscillator and still the hardware of the digital correlador. All the devices will be applicated in a radiometer / polarimeter for detention of polarized radiation at microwaves, coming from the outer space.

1.4. Methodology

The work developed that gave rise to this thesis is structuralized in some stages:

1. Theoretical study of some inquiries related with galactic emission mapping that served as introduction to the RA and current knowledge of some existing projects. Information for this study had been supplied by elements of GEM. In this point it is intended to describe the complete project in a block diagram, defining the function to execute for each block.

2. Theoretical study RF Electronics and Microwaves. The literature used for this study consisted of notes and books from lectures of RF Electronics. The studied literature involved more specifically theory of filters, low noise amplifiers, mixers

and oscillators. All the characteristics (compression point, intermodulation distortion, dynamic range, sensitivity) involved in the implementation of such devices were studied in more detail

3. Study of dedicated electronic design and RF electronic/microwave simulation software. For the effect ADS (Advanced Design System) was used to simulate the RF and microwave circuits. The schematic and layout for all the circuits present in the system were designed using ORCAD 9.1.

4. Design of the schematic, simulation and designing the layout for all the circuits developed throughout this work, like: an IF (Intermediate Frequency) pre amplifier with a filtering factor, an IF amplifier with digital control attenuation, a converter of frequency from IF to base band, a 600 MHz local oscillator and microwave filter with 700MHz bandwidth centered at 5GHz.

5. Test and measurement results of all the printed circuit boards (PCB). The analysis of the results will be able to complete the desired function of each circuit, being able to modify the circuit or its components, to attain the desired results.

1.5. Implemented Circuits

An IF chain that will integrate a radio telescope, currently developed for GEM-P project, was implemented, this chain is composed by:

1. IF pre amplifier – it has a set of two amplifiers stages and a band pass filter. This circuit besides amplifying it also performs a selection of the IF frequency band of the entire system. It uses the latest RF technology and performs protection from interferences from external sources.

2. IF Amplifier – this circuit is the largest gain contributor of the system, it has a set of five amplifying stages together with two digitally control attenuators. Like the previous circuit also uses the latest RF technology and high level of protection.

3. Converter – it performs the second down conversion of the system, this time from IF (600 MHz) to zero IF (base band). The conversion modulates the signal in (I) phase and (Q) Quadrature, it also performs a small amount of voltage gain, to feed the ADCs with the desired level of signal. It also uses the latest RF technology.

4. Local Oscillator – It provides the mixers from the converter with the necessary signal to multiply with IF at a frequency equal to the center frequency in IF. The power is defined by the LO port from the mixers. To synthesize the frequency it has a PLL.

In order to perform the first selectivity step of the system, it was also implemented a microwave band pass filter, made with passive components, using microwave technology. It is a coupled line filter with microstrip transmission lines. This circuit avoids that undesired signals enter the mixer.

1.6. Thesis structure

This thesis is organized in several chapters:

- The second chapter describes the entire receiver to be developed for the GEM-P project. It explains how a receiver works and the types of receivers that exist. A more detailed description of each block developed, is also defined in this chapter as also the problems encountered and the ways taken to solve it. The measured and tested results are also shown.

- The third chapter details the converter circuit and design constraints, it is not included in chapter two, because it is a new application in this type of receivers. A description of the local oscillator implementation is also present in this chapter.

- The chapter four is the hardware design of the full digital correlator and a brief description of its functioning. It is composed of four ADC that digitize the signal to feed a FPGA that will make all the calculations needed to determinate the Stokes Parameters

1.7. Original Publications

Throughout the work carried through for this thesis two articles in international conferences had been published and still an oral presentation and a poster in one Radio astronomy workshop. The articles in question focused diverse aspects related with the set of IF circuits and also it served to give a brief description of GEM-P project the present public, nominated:

- Workshop Digital Receivers – RADIONET, Bolonha, Abril de 2007 – “GEM-P – an FPGA based Polarimeter”
- Conftele2007, Peniche, Maio de 2007 – “Design of an IF Section for a Galactic Emission Mapping experiment”

2. Receiver

Basically a radiometer is a calibrated, high sensitivity microwave receiver, with the function of measuring and detect celestial emission (a radiometer can be used in other ways, but it follows the same structure, in this case it is a receiver to apply in a radio telescope). Many times this type of emission is not so different from the noise generated by the own receiver or even from backend radiation coupled to the receiver. Usually the signal level rounds 10-15 to 10-20 Watts. So it is extremely necessary the implementation of a receiver very well calibrated and with high sensitivity.

Theoretically speaking a radiometer facilitates the measurement of a brightness temperature object. For that is used an idealized antenna pointed towards the object and the emitted power (corresponding to the brightness temperature of the object) is collected by the antenna. At the output, in the case of a lossless antenna, it will be an output power TA that is directed related with the brightness temperature of the object. The task of the microwave receiver is to measure this temperature with sufficient resolution and accuracy.

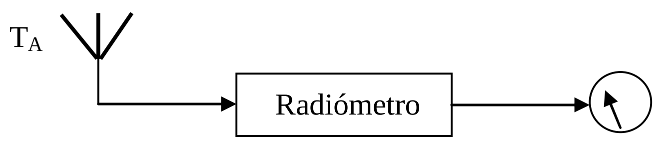

Figure 8 – Ideal radiometer

The radiometer selects a portion of the available output power from the antenna, that is, a certain bandwidth B around a given centre frequency. This power is amplified (G) and outputted to a medium, correlator, power meter. The meter measures:

$$\begin{aligned} &P = KBGT_A\, Watts, \\ &K - \text{Constante de Boltzmann} = 1{,}38 \times 10^{-23}\ J/K \end{aligned} \tag{1}$$

In a real environment a radiometer generates noise and this noise will add to the input signal

$$P = KBG(T_A + T_N)\, Watts \tag{2}$$

Where T_N is the noise temperature introduced by the receiver. To all the radiometers is associated a sensitivity problem, that con be described by the resulting formula that corresponds to the standard deviation of the output signal.

$$\Delta T = \frac{T_A + T_N}{\sqrt{B \cdot \tau}} \tag{3}$$

This is the basic radiometer sensitivity formula, in which TA is the input temperature to the radiometer, TN its noise temperature, B its bandwidth and ζ its integration time. The accuracy is also an important performance and is dependent of gain and noise temperature caused by active components, like amplifiers, that are dependent on supply voltage.

The prime tasks of a radiometer are input frequency band selection and amplification of the incoming signal to a proper level for backend circuitry. The radiometer structure is basically an amplifier followed by a filter that selects the desired frequency portion and finally a frequency conversion. The next components have the basic IF characteristics with an amplification and filtering once again. The Friis equation tells that the first component is the greatest contributor of noise.

$$NF = NF_1 + \frac{NF_2 - 1}{G_1} + \frac{NF_2 - 1}{G_1 G_2} + ... + \frac{NF_n - 1}{G_1 G_2 ... G_{n-1}} \quad (4)$$

This formula means that the first component in this chain is the strongest contributor of the cascaded Noise Figure of the entire system, this way the first amplifier will be a Super Low Noise Amplifier (LNA). Its behavior is mainly to have the lowest NF (below 0,3 dB). The frequency conversion is made by a mixer and a local oscillator (LO). Mixer is an element that at a determined frequency becomes non-linear allowing for signal multiplication with different frequencies. By this component it is possible to go down and up in frequency. The multiplier signal is provided by the LO, that exhibits a fixed frequency, equal to the difference of frequency needed.

In order to avoid the accuracy degradation there are principles that can be used to surpass such problems, in this report it will only be referred the Dicke Radiometer (DR), Noise Injection Radiometer (NIR) and Total Power Radiometer (TPR). The last is described by a block diagram in figure 9.

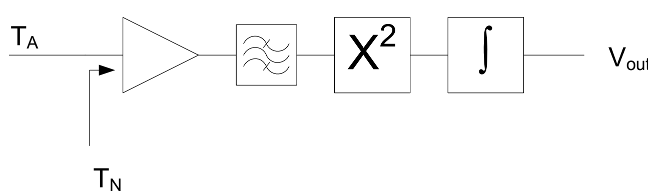

Figure 9 – Total Power Block Diagram

An amplifier with gain G symbolizes the gain of the radiometer, the frequency selectivity is defined by a filter with bandwidth B centered on a desired frequency (for this work it will be around 4.9 GHz). Next is the square law detector to measure the signal mean and finally an integrator to reduce output fluctuations from the detector. At the output it will be present:

$$V_{out} = c \cdot (T_A + T_N) \cdot G \tag{5}$$

where c is a constant. V_{out} is totally dependent on T_N and G. The TPR sensitivity is equal to (3). DR does not measure directly the antenna temperature, instead it switches between the antenna temperature and some known reference temperature, at the output will be the difference between these two temperatures, as can be verified in equation (6).

$$Vout = c \cdot (T_A - T_R) \cdot G \tag{6}$$

The sensitivity is greatly reduced since in this topology the noise temperature and gain fluctuations are also decreased. The sensitivity formula for DR is in equation (7).

$$\Delta T = 2 \cdot \frac{T_A + T_N}{\sqrt{B \cdot \tau}} \tag{7}$$

The NIR is an improvement of the DR, the output is independent of gain and noise temperature fluctuations.

$$\begin{aligned} Vout &= c \cdot (T_A{'} - T_R) \cdot G \\ T_A{'} &= T_A + T_I \\ T_A &= T_R - T_I \end{aligned} \tag{8}$$

The sensitivity is similar to that of the DR:

$$\Delta T = 2 \cdot \frac{T_A{'} + T_N}{\sqrt{B \cdot \tau}} \tag{9}$$

A radiometer is essentially a transducer that is responsible to translate the signal gathered by an antenna and transform it in a way that allows a acquisition system to understand it and make the necessary calculations. Its front-end circuitry is divided in two prime tasks: input frequency band selection and amplification of the level of the input signal to a proper level in order to be handled by the low-end circuitry. The amplification is normally very large, typically 60-80 dB for microwave radiometers, and it can be implemented in two different ways: direct receiver and superheterodyne receiver. In the direct receiver all the amplification and selectivity takes place at the input frequency (RF range), on the other hand, for the superheterodyne the amplification is defined at a much lower frequency (IF) and the selectivity is a combination of filters at RF and IF.

The amplification could also be a combination of RF and IF amplifiers, usually at RF the gains rounds 10-30 dB. The first selectivity is applied in RF with a filter having a larger bandwidth than at IF, where the selectivity takes place. The mixer “brings” down the frequency with the help of a strong signal provided by a local oscillator (LO), producing at the output a signal at IF that is proportional to the power of the RF input signal, the final selectivity is accomplished by a filter at IF. Once again the signal amplified but with a greater gain than at RF, between 60-90 dB that is going to feed with the needed value a detector, an integrator, a data acquisition system that can be digital or not. In the GEM-P case, the radiometer used will be a Superheterodyne Noise Injection Radiometer with double down frequency conversion. Another feature in addition to determine the signal power will be the calculation of the input signal polarization, this way this receiver is a radiometer / polarimeter.

The correlation, integration and data analysis will be executed totally in digital domain, being a new and pioneer approach to this technique in radiometry. Digital correlation in polarimetry is based on the cross correlation of the right and left circular polarizations as seen by the Stokes Parameters. In the digital domain there is the advantage of avoiding mixing signals once it are digitized, besides it is of easy implementation.

Another important characteristic is the sensitivity that will be less than 1 mK (Kelvin), an Instantaneous Dynamic Range of 20 dB and a Total Dynamic Range of 80 dB. In figure 10 is presented the Block Diagram in which is described the several elements of the radiometer / polarimeter being developed for GEM-P.

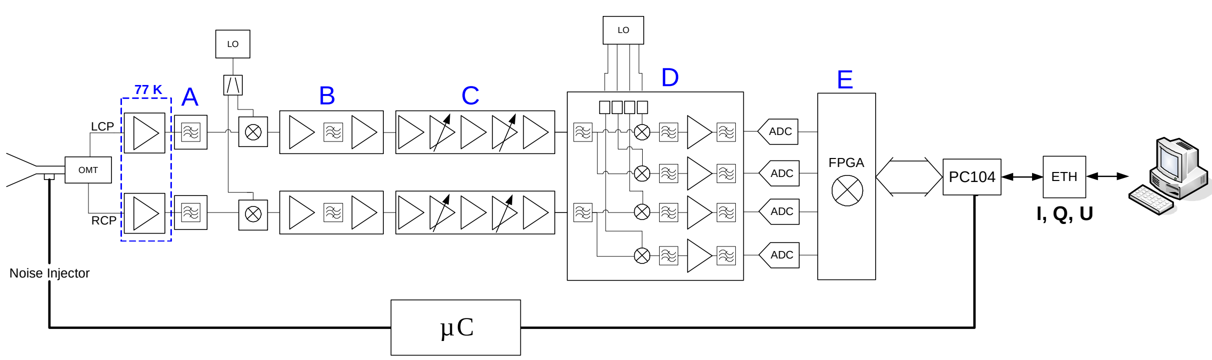

Figure 10 – Receiver Block Diagram

The structure of a Radio Astronomy (RA) receiver is identical to that of a Telecommunication Receiver, like it can be verified in the Diagram presented in Figure 10. But there are some differences in the characteristics of each block, more precisely, in the kind of information of data that the receiver deals. Unlike telecommunications, where receiver performance is described in power units - dB, in RA it refers temperature units - Kelvin (or temperature units). By definition, in RA the incoming radiation of the sky is expressed as sky temperature radiation. The signal in RA is much lower than that for Telecommunication (for GEM-P it is below 1 mK) meaning that the instantaneous received signal has usually a much smaller magnitude than noise. The frequency bandwidth in Telecommunication receivers are also lower than that for RA, a normal bandwidth is 10 MHz or less, In RA bandwidth is usually up to 10% or more of then central bandwidth. So, there is the problem of spreading of noise in a greater band.. In order to obtain the desired results the signal is correlated and integrated, this way avoiding the error associated with small LNA gain fluctuations.

Thus, to guarantee a reception without bit errors, fluctuations introduced by the receiver need to be avoided at all cost. This can be achieved providing a good flatness of gain over the required bandwidth. This is easy accomplished in small frequency bandwidths, like in Telecommunications receivers. Usually oscillations of the order of 1 dB are acceptable, but for RA are disastrous (less than 0,1 dB for GEM-P). The gain flatness allows that all the data in the frequency band is amplified at the same way, guaranteeing a reception with no errors.

The blocks symbolized by letters, in figure 10, are already designed, simulated and tested. The LNA is being developed.

This receiver as the basic superheterodyne topology, the back-end being fully implemented in digital domain (Figure 1). In radiometry the bandwidth should be as large as possible to permit the best instrument resolution. This is however limited by the available bandwidth free of interference and preferably under protection of the international frequency allocations for the RA (radio-astronomy) service. In this project it is wanted a minimum bandwidth of 200MHz around 5GHz. However, the center frequency had to be changed to 4.9GHz (using the same 200MHz bandwidth) in order to be aligned with the band segment allocated to RA for which it can be applied for protection of the Portuguese radio spectrum administration

(ANACOM). The radiometer / polarimeter receiver is a double conversion super-heterodyne receiver with zero-IF. The front-end will use cryogenically cooled HEMT preamplifiers followed by image rejection filter and diode mixers along with a local oscillator. All this equipment will be located at the back end of the antenna feed inside a temperature shield. The IF chain is composed of a preamplifier-filter and a large gain IF amplifier followed by a converter to zero-IF that provides the base-band signals for the correlator.

The first block is the antenna, for calibration purposes is used a noise injector in conjunction with the antenna set up, followed by the OMT (OrthoMode Transducer) that separates in left and right circular polarization. To improve sensitivity LNA (Low Noise Amplifiers are used to amplify the signal, thus contributing to noise reduction of the receiver, it also amplifies the signal of about 30 dB. The signal is than filtered by a image rejections filter, having a bandwidth of 600MHz at center frequency 4,9 GHz. The first down conversion is accomplished using one mixer and by a strong signal from a commercially local oscillator that converts to IF or 600 MHz. The following two blocks (B and C) follow the classical IF characteristics, the first is the IF pre amplifier filter, it amplifies 31 dB and restricts the bandwidth of the receiver to 200 MHz at 600 MHz. The other IF block is the IF Amplifier, it contributes with the largest amount of gain in the system – 71 dB, it also has built-in gain adjustment capabilities using digital control attenuation. Next is the second down conversion from 600MHz to zero IF, using Phase and Quadrature modulation scheme, by a strong signal from Local Oscillator that feeds the Converter with four signals, with 90º phase difference. Before Analog to Digital Conversion in the Correlator the signal is again voltage amplified. The four ADCs digitize the signal to feed a FPGA that is responsible to correlate and Integrate the signal, the FPGA also sends these data to a Control PC (PC104) via ISA Bus. The PC104 also integrates and create the files that describe the information of the radiation gathered by the antenna. The files are then sent to a PC elsewhere using Ethernet to analysis. The Microcontroller is there to gather information and control some procedures of the environment of the antenna, namely, the Noise Injector, time, temperature, wind speed, rain, position of the antenna - zenith and elevation, motion of the antenna, reset.

The following table shows the gain and attenuation distribution along the receiver stages that describes the power gain budget:

Antenna	**LNA**	**Passive Filter**	**Mixer**	**IF Pre Amplifier**	**IF Amplifier**	**Converter**	ADC
Input (dBm)	26	-4	-7	31	56	2	Output(dBm)
-105,6	-79,6	-83,6	-90,6	-59,6	-3,6	-1,6	-2

Table 1 – Power gain budget of the receiver

The input is referred to the signal from the antenna feeding the OMT and the output symbolizes the maximum level signal accepted by the ADC, respectively -105,6 dBm and -2 dBm. The values represent assumptions made for each stage, for example, the first mixer normally has a conversion loss near -7 dB, the passive filter has an attenuation near -4 dB, like it will be seen, these values are very close to the real, has it will be shown. The attenuation values are known, the gain factors can be easily distributed accordingly by the rest of the amplifying stages, it were attributed the gain factors by the rest of the blocks. As was said before, at RF the amplification varies between 10 and 30 dB, the rest of the gain will be in IF. So the total gain of the receiver is 104dB, divided in four blocks: RF, two IF amplifiers and signal amplification in base-band in the converter. During this thesis are only referred some blocks of the receiver, pointed out by letters. They are the B - IF Pre Amplifier Filter, C - IF Amplifier, D – Converter and Local Oscillator, E - Digital Correlator and also at RF the A - Image Rejection Filter, the LNA in a development phase.

This chapter describes the system requirements and derivation of the components specification of each block of the system, it also presents design options and constraints presented along with the simulations and experimental results, except for LNA, which is in a developing phase.

2.1. RF

This superheterodine receiver is especially designed to fulfill standard characteristics of a RA experiment. In this sub chapter will be specified the system requirements of the RF part, which determinates the total sensitivity of the system and also requires less gain fluctuations, it also filters undesired signals from outside the RF band of interest (selectivity). Since the

LNA is not implemented yet it only refers the design options and some simulations results obtained to accomplish the required gain, noise and stability.

2.1.1. Image Rejection Filter

As the frequencies reach the region where lumped elements cannot be practically realized, there is a need to build filters with transmission line components. Much of the theory used for low frequency filters is also applicable to microwave filters except different elements are used to realize the filters. Inductors are replaced with short circuited transmission line stubs and capacitors with open circuited transmission line stubs. Periodic structures generally exhibit pass band and stop band characteristics in various bands of wave number determined by the nature of the structure. When two unshielded transmission lines are close together power can be coupled between the lines due to the interaction of the electromagnetic fields of each line, such lines are referred to as coupled transmission lines and usually consist of three or more conductors in close proximity. The coupled lines can be represented by the structure shown in the next figure:

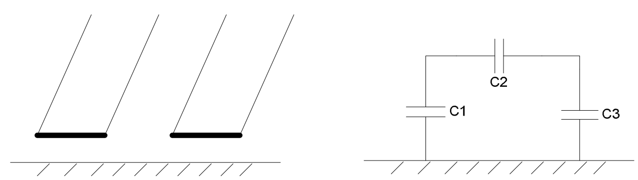

Figure 11 – Coupled Lines equivalent circuit

C1 and C3 represent the capacitance between one strip conductor and ground, C2 represents the capacitance between the two strip conductors. This type of lines can be used to construct many types of filters. Coupled transmission lines have frequency sensitive coupling, and can be analyzed by the even-odd mode method. In particular, the configuration that represents coupled λ/2 open lines is the easiest to construct in microstrip and strip line. Fabrication of multisection band pass coupled line filters is particularly easy in microstrip for bandwidths less than 20%. Wider bandwidth requires very tightly coupled lines, which are difficult to fabricate.

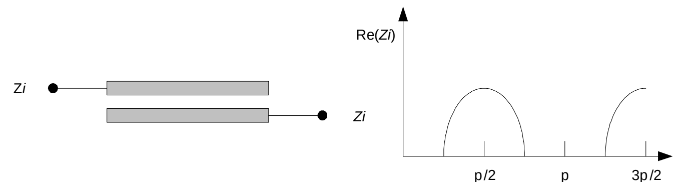

Figure 12 – Coupled Lines frequency Response

So it can be seen that a structure of a number of coupled lines will admit to an equivalent circuit of alternating series and parallel resonant circuits There are other combinations of terminating the four ports [7] of this coupled line section, but the interest is in creating a band pass filter and open circuits are easier to fabricate than are short circuits. The purpose of this filter is to eliminate unwanted signals lying outside the RF band containing the possible information to be detected. Unwanted signals can include signals fed from the antenna, and due to gain roll off of the preceding amplifier. The LNA will provide gain to all frequencies within the RF bandwidth and its gain is likely to roll off beyond it. Furthermore, the amplifier will amplify noise across the entire band, and possibly at the image frequency as well. Therefore this component suppresses undesired signals in particular the image frequency maintaining the system NF by preventing image noise from entering the mixer.

To design the filter was used ADS 2005 (Advanced Design System) from Agilent, since this component is based in electromagnetic procedures, ADS is the best solution, which combines circuit and electromagnetic simulation. The filter will be centered at 4,9 GHz having 600MHz bandwidth. The order selected a priori was 4, meaning that it will be four sections identical to the one in figure 13.

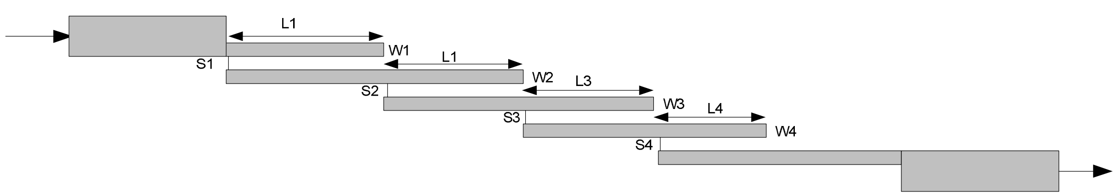

Figure 14 – Microwave image rejection Filter

The lines at the beginning and at the end represent 50Ω lines to connect to outside cables. To define the function of this section there are several variables to know, namely Wi, Li and Si, i=1, 2, 3, 4. Wi represents the with of both lines of each section, Li is the length of each section and Si is the space between lines of each section. But ADS has a facility to use while designing passive circuit, which is the case. The Design Guide (DG) creates the circuit according to the desired stop/pass band frequency, stop/pass band, characteristic impedance, response type and order values introduced by the user. There are other specifications, but these are the important ones. The other essential characteristic to develop a microstrip filter in ADS is to select the substrate. A small PCB board was donated to me to make the filter, it has a RO4003C substrate from Rogers and it offer superior high frequency performance and low cost circuit fabrication. RO material possesses the properties needed by RF/microwave circuit designers. Stable dielectric properties over environmental conditions allow for filter design. The low dielectric loss allows the use at higher frequencies than conventional circuit boards. The type of signal to be detected is very low, in the order of sub miliKelvin, this laminate is ideal for sensitive temperature applications. The 5GHz frequency is not also a problem because the dielectric constant is very stable over a broad frequency range (10 GHz). The more important characteristic to include in the substrate definition in ADS are:

Substrate Thickness	H	20 mil
Relative Dielectric Constant	ε_r	3,38
Conductor Thickness	T	0,35 µm
Dielectric Loss Tangent	tan δ	0,0021

Table 2 – Substrate Values

To create a filter with these characteristics using Design Guide was inserted a Smart Component for the type of filter wanted, in this case is a Coupled Line Filter. The next step is to introduce the stop/pass band frequency, stop/pass band attenuation and order values. The substrate “Msub1” in the Smart Component represents the Substrate used.

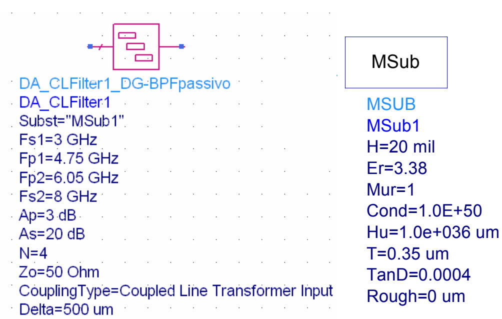

Figure 13 – Passive Filter Characteristics

The values for the pass band frequency are strange, because they are the final values obtained after several adjustments made to guarantee a band of 600/700MHz around 4,9GHz, and since the Design Guide (DG) did not retrieve the necessary filter, therefore it were made several simulations till the good results arose. The coupled line filter created is presented in the figure below:

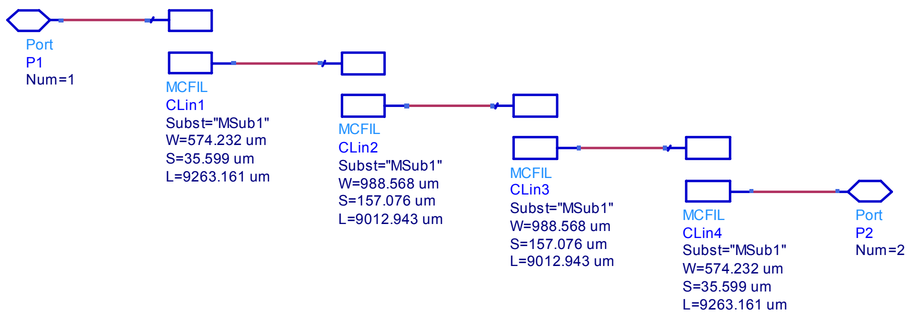

Figure 14 – Filter Schematic from ADS

The DG designs a filter determining the width (W), length (L) and spacing between lines (S) for each section of the filter. Analyzing the values it emphasizes that it is a symmetrical filter.

To show what this circuit gives rise, it was done an S – Parameter simulation, from DC to 20 GHz. The results for a 50Ω input/output line are shown in the figure below:

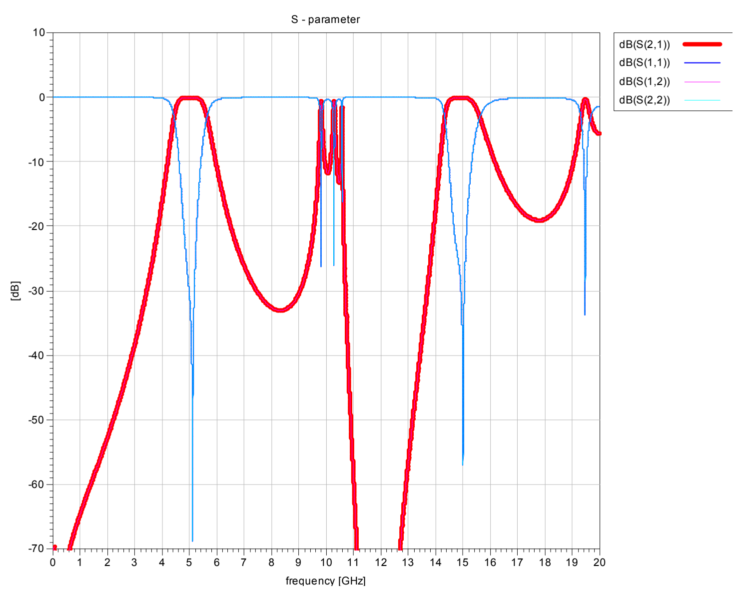

Figure 15 – Simulated S parameters of Microwave Filter

S-parameter simulation calculates S11, S12, S21 and S22 values between the ports of the filter. The graphic above shows the four S-parameters, S11 is equal to S22 and S21 is equal to S12. S11 (S22) in blue, verifies that the input (output) is matched around 5GHz and S21 (S12), in red, means that the response of the filter is very well defined, having a bandwidth of ≈750MHz centered at 4,9GHz, the gain is maintained flat along the pass band with minimum losses (-0,1 dB). The harmonics of the fundamental are also present, some having unity gain, namely at 10GHz and 15GHz, these bands need to be attenuated as possible to avoid their presence at the input of the mixer, next to the image rejection filter, the way to solve this problem, will be explained further. Another tool of ADS enables the generation of the layout:

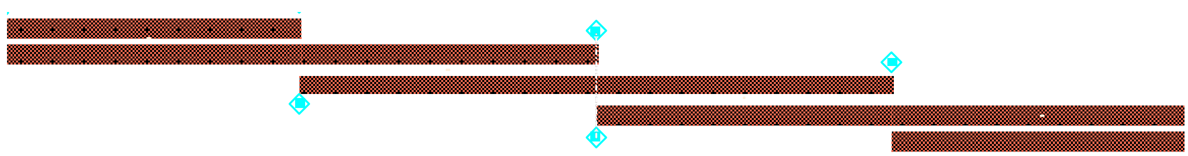

Figure 16 – Layout of Microwave filter

$$\boxed{\begin{aligned} f_{IF} &= f_{RF} - f_{LO} \\ &= f_{LO} - f_{RF} \end{aligned}} \quad (10)$$

The mixer translates all the incoming signals in the RF frequency range into signals in IF, basically it down converts from 4,9GHz to 600MHz, so the strong signal generated by the Local Oscillator is tuned at the 4,3GHz. The IF result it will be a bandwidth of 1GHz around 600MHz. The problem is that are weaker harmonics of the LO strong signal, that also feed the mixer and may convert undesired bands causing IF degradation. These formulas help to explain the reason of the problems. What is needed is the fundamental conversion, that is, f_{LO}=4,3GHz, f_{RF}=4,9GHz and f_{IF}=600MHz, but the filter response shows spurious at 9 to 10GHz. LO harmonics (8,6GHz, 11,9GHz …) feeding the mixer, can cause IF at 400 MHz to 1,4GHz, meaning that the spurious lie down in the desired IF band, destroying the original signal. The solution was introducing stubs at the input and output to attenuate the signals by loading the circuit at the undesired frequencies. At higher frequencies is difficult to implement short circuited stubs, so the solution was open circuit stubs. To link the stubs to the circuit were introduced a 50 Ω lines between both sides of the stub. The circuit with the stubs is presented in the figure 19:

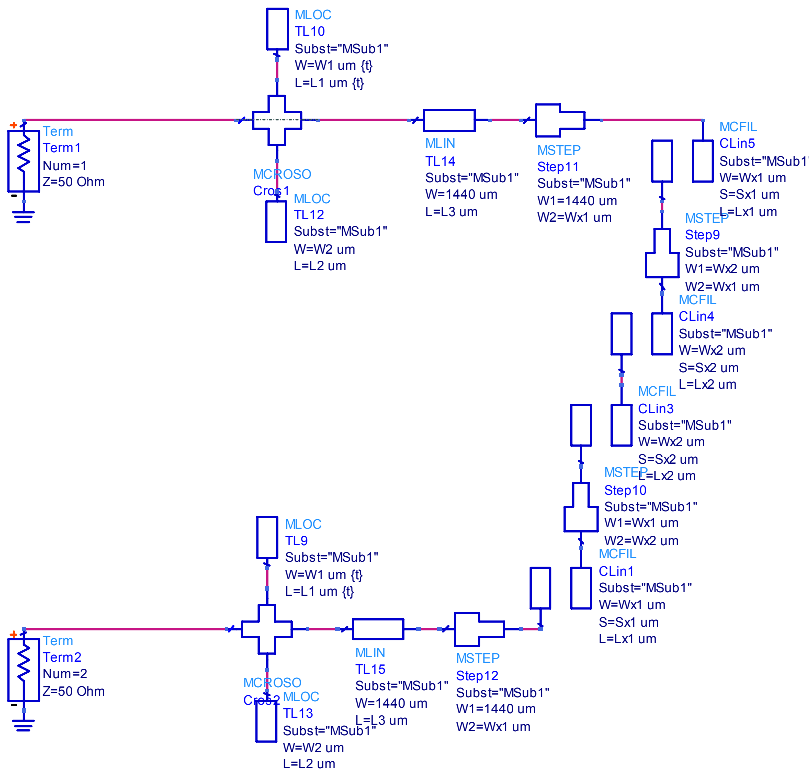

Figure 17 – Final schematic of Microwave filter after adjustments

The circuit in figure 19 already has the steps between lines with different widths; the first circuit suffers from discontinuities because of this missing component. As you see there are several values not specified, instead are letters, the reason for this, is because the results weren't so good and to adjust the characteristics of each component the values of Width, Length and Spacing between Lines were varied to achieve the desired function for our filter. It was found that the circuit did not retrieve good results because of the lack of steps between transmission limes that caused bad connections that were visible in the S parameters Simulation. Still is missing the 50 Ω transmission line responsible to guarantee a match at the input and output of the circuit. The trial and error simulation gave, after several simulations calibrating and adjusting, to the result in the graphic of figure 20.

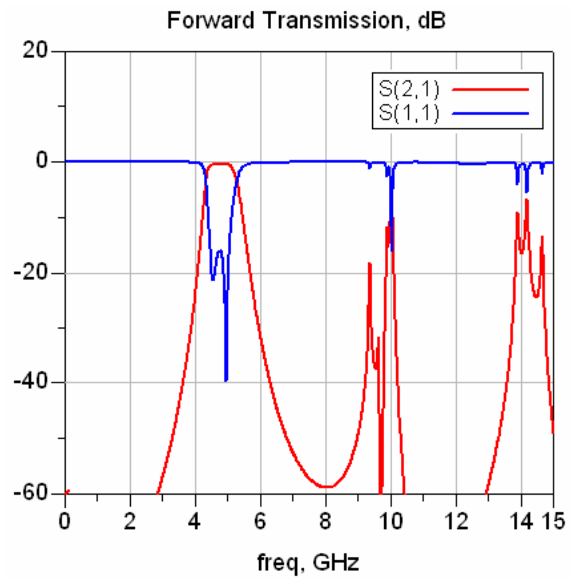

Figure 18 – Final simulation results of microwave filter

The unwanted frequencies are attenuated at least 20dB, maintaining the right filtering. At 10GHz is less attenuated but will result in a 1,4GHz IF after down converting and does not affect the bandwidth of the IF chain. These are the results using ideal components, what is needed is the result closest to the reality and ADS has the right tool for that, is called Momentum. Momentum is an electromagnetic (EM) simulator that computes S-parameters for general planar circuits, including microstrip, slot line, strip line, coplanar waveguide, and other topologies. Also gives a complete tool set to predict the performance of high-frequency circuit boards. For this specific case it is very useful identifying parasitic coupling between components. Accurate EM simulation improves passive circuit performance and increases confidence that the manufactured product will function as simulated. The layout is generated automatically with the help of the Generate/Update tool from ADS,

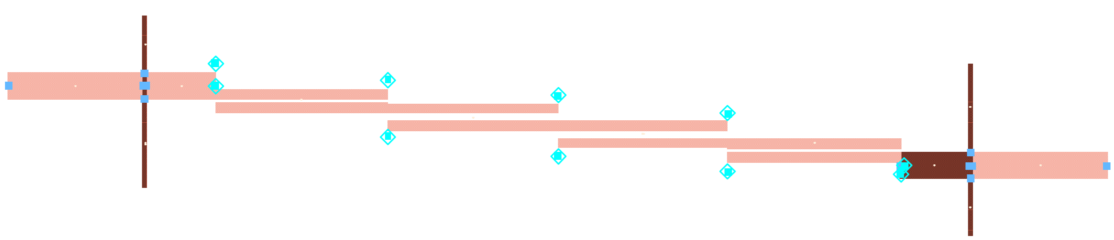

Figure 19 – Final layout of filter

The larger lines at the input and output where the stubs are connected are 50 Ω lines, and are useful to link the SMA connectors to the Printed Circuit Board (PCB). The EM simulation presents very satisfactory results, very near to those from the circuit simulation.

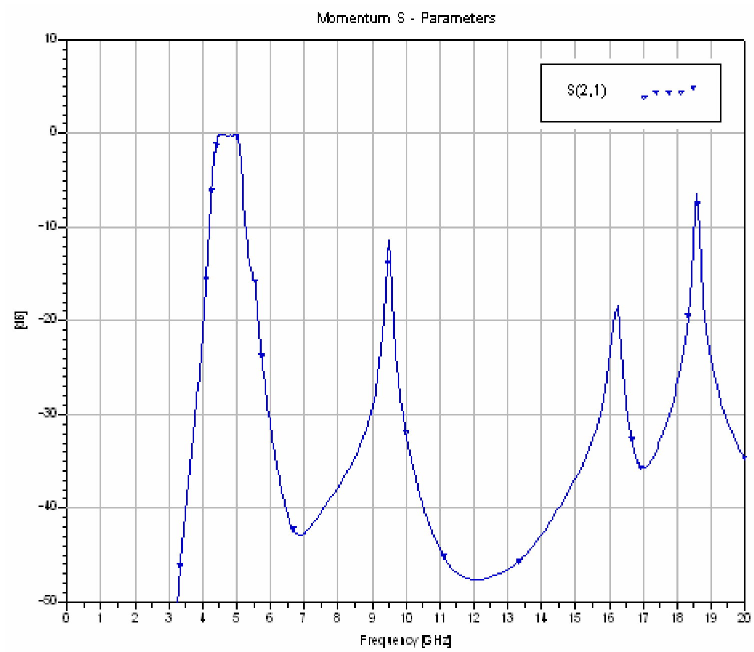

Figure 20 – Electromagnetic simulation (Momentum) result of passive filter

In figure 22 is the S21 parameter obtained from Momentum simulation. It is visible that is preserved 1GHz around 4,9GHz and the undesired frequencies are attenuated. Now, having found the response needed it is time to design the layout. Using ADS the layout was exported to edit it in AUTOCAD, editing was for redrawing the layout to fit in a rectangular box, that was done rotating the layout about 5º and curving the input and output lines, keeping the initial geometry. These changes gave rise to the final layout:

Figure 21 – Final layout of passive filter

The dimensions are 60 × 35 mm. Small variations of the length, width or space between lines sizes can damage the response, the implementation of the PCB board was very careful. A photo-lithographical process was used to guarantee a better resolution of the final printed circuit, since direct printing (CNC and others) does not have such high resolution. The PCB board was packaged in an aluminum milled box specially designed for this type of application to serve as shielding.

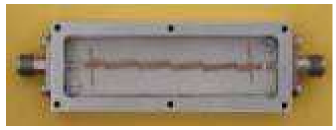

Figure 22 – Photo of filter already in the alumina box

A network analyzer from Agilent E8361A worked for testing the filter. Once again the S parameters returned the behavior of this circuit, like are presented below:

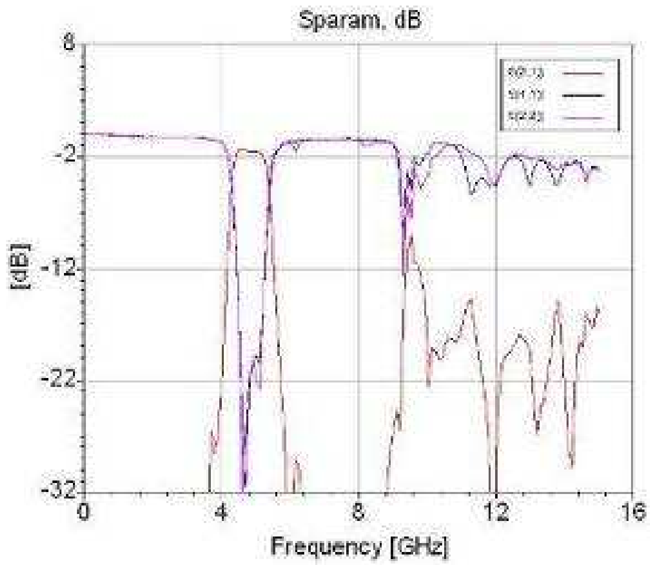

Figure 23 – Tests results of filter in Agilent E8361A

As shown in the figure above, the filter presents 1GHz flat band around 4,9GHz with good cut-off frequency definition as well as the attenuation outside the desired band. As obvious it presents matched response at input and output in the band, displayed by S11 and S22 parameters. In resume, the simulations are very near the tested results.

2.2. Intermediate frequency

Receiver for radio astronomy experiments are similar in construction to receivers used in other branches of radio science and engineering. Basically the IF chain for GEM-P follows the classical characteristics of IF receivers. As was said before, this is a super heterodyne (SH) receiver, the signal is coupled to the receiver by an antenna, near the feed is the front-end block where the signal is amplified at RF (LNA), RF preselected (Image Rejection Filter) and a mixer to convert to a lower intermediate frequency, in this case is 600MHz. The signals coming from the front-end block enter the IF unit of the receiver. In a SH receiver the largest part of the gain is obtained in IF and also determines the receiver bandwidth. The IF chain is divided in two parts: an IF pre amplifier – filter and an IF amplifier. The first circuit is a combination of an IF pre amplifier and an IF filter. The IF preamplifier provides adequate gain to drive the following stages, the IF filter is a band pass to reject the unwanted signals generated by the mixer and other components, also removes any DC offset and out of band frequencies. The IF Amplifier has the largest gain contribution of the receiver, it also performs digitally controlled attenuation to manage the signal level at the input of the ADCs.

In this sub chapter is divided in two parts: one for the IF Preamplifier – Filter and other for the IF Amplifier.

2.2.1. IF PreAmplifier – Filter

The signals coming from the front-end block (located at the feed point) enter the IF unit of the receiver directly to the IF preamplifier – filter module, the kind of data that is going to be extracted from the received signals are in the order sub-miliKelvin of antenna temperature which are related to the receiver gain stability. It is essential to ensure that gain variations would preferably be in a different time scale and smaller than the measured variations. The variation of gain with environmental temperature needs to be minimized at all cost. Therefore amplifiers with minimum gain variation with temperature were selected. The IF filter presented in this circuit is high Q to prevent oscillations in the pass band and define as well the cutoff frequency and attenuation. The investigation started by creating a filter with the mentioned characteristics. A Butterworth type filter seemed to be an attractive choice because of its flatness. The elements are lumped components placed in a T configuration. Basically it is a pass band filter centered at 600MHz with 200MHz bandwidth. The quality factor (Q) represents the sharpness of the filter, or rate that the amplitude falls as the input frequency moves away from the centre frequency. To achieve high Q values the inductances were hand made using silver with air nucleus. This procedure guarantees a Q factor of 300, much higher when compared to commercial inductances (75). To determinate the number of turns, coil turn radius and coil turn length, was used the characteristically formula:

$$L(\mu H) = \frac{0,363a^2N^2}{9a+10b} \qquad (12)$$

a(cm) –coil turn radii
b(cm) – coil turn length
N – Number of turns

For the 3 nH inductor results → a = 1 mm; b = 4 mm and N = 2 turns,

For the 27,2 nH inductor results → a = 2 mm; b = 3 mm and N = 3 turns.

The inductors were created using typical wired wrap.

The defined characteristics for the filter gave rise to a filter identical to the one in figure 26:

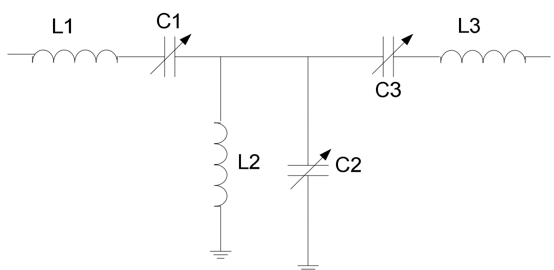

Figure 24 – Band pass filter schematic preview

The design of the filter as well all the other parts was computer aided, using the Advanced Design System (ADS) software. An ADS Tool guides the implementation of the filter without analytical calculations just by introducing the claimed values. The first results gave to low capacitor (850fF) and high inductor (70 nH) values, that needed to be changed. A fine adjustment was made in the final design by trial end error, in order to achieve the desired bandwidth and response flatness with commercially available component values. After several regulations it produced the following simulated S parameters:

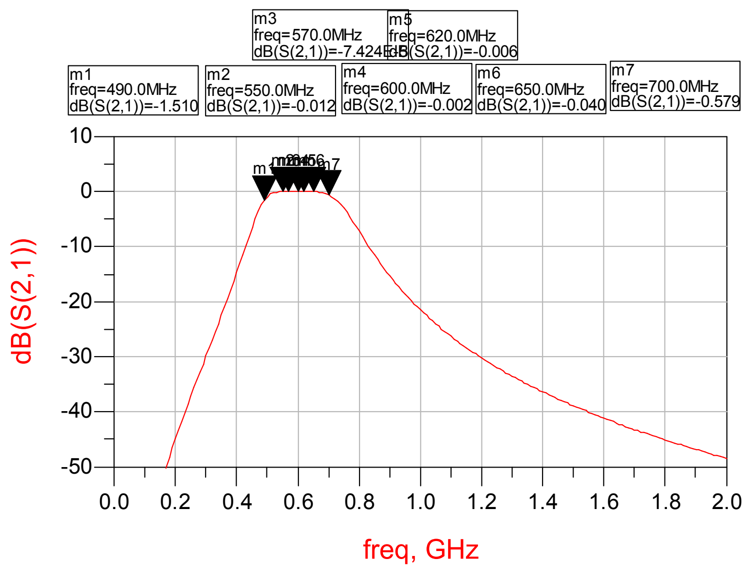

Figure 25 – S21 parameter simulation of band pass filter

This graph meets all the needs but fabrication in PCB could be a consequence of some displacement of the frequency. To allow calibration the filter uses trimmer capacitors and fixed inductors.

This module needs to present a moderately low Noise Figure (NF), since the filter has high NF was placed between the two amplifiers (with lower NF) of this module improving the NF. This module presents a gain of 31 dB its design was targeted for gain and proper IF bandwidth shaping purposes but other considerations were taken into account, such as gain variation with temperature and frequency. The variation of gain with environmental temperature needs to be minimized at all cost. Therefore amplifiers with minimum gain and variation with temperature and frequency were selected. Encapsulated MMIC (monolithic microwave integrated circuits) were attractive choices and their possible use was investigated. Very wideband MMIC's do not exhibit a constant gain over frequency, higher frequencies presenting the lowest gain. In order to reduce this effect it was inserted a slope compensation network, which is a simple RLC network, next to the filter that reduces the gain at lower frequencies by loading the circuit more heavily at lower frequencies.

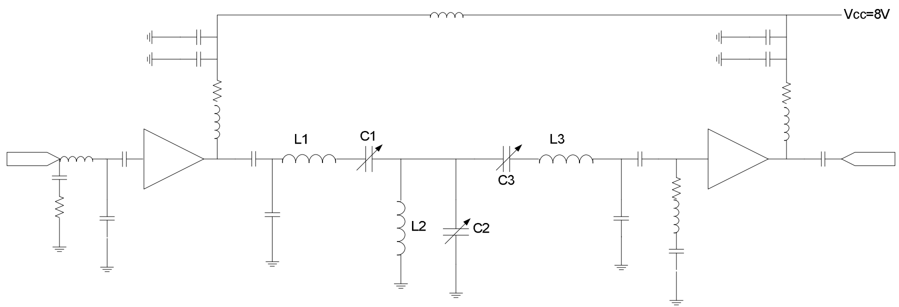

Figure 26 – IF pre amplifier Schematic

As shown in Figure 28 this module has two amplifier stages with a band-pass filter between them. The preceding circuit is a mixer that provides the first down conversion of the system, It has been shown that the intermodulation performance of mixers can be improved if care is taken to properly terminate undesired leakage signals and mixing products. One way to do this is through the use of diplexing filters. A diplexer is basically a frequency multiplexer that splits a single channel carrying many frequencies into two channels carrying fewer frequencies. In the design presented here, frequency selectivity is accomplished by placing a low pass filter in parallel with a high pass filter. The diplexer sees 50 ohm impedance looking into the amplifier and back at the mixer. The IF signal to the mixer is at 600 MHz, with signal bandwidth of 200 MHz. The mixer LO frequency is at 4,4 MHz. All of these signals enter the diplexer where they are split. The desired signal is passed through the amplifier channel to the low pass. The diplexer is formed by paralleling singly terminated low and high pass filters derived from the same normalized low pass prototype. The undesired signals are passed through the high pass channel and are dissipated in the 50 ohm resistor. Terminating the undesired signals in this manner minimizes reflections back into the mixer where further harmonic generation can occur.

DC Block capacitors were placed in the circuit to prevent DC signals. The amplifiers used were ERA2 and ERA3 (from Minicircuits) due to their small temperature drift in the frequency band of interest (500MHz to 700MHz). The Bias resistance from the first amplifier (ERA3) is higher than the second one (ERA2) in order to improve noise in the first ERA and improve gain in the second ERA. The manufacturer claims a gain increase with temperature of 0.12 dB from -45 to 85ºC, which corresponds to 0.0009dB/ºC. Taking into account that the temperature of the whole system will be controlled to 1ºC, this is enough for this application. On the modeling was used ERA3 and ERA2 and the results are shown in Figure 29

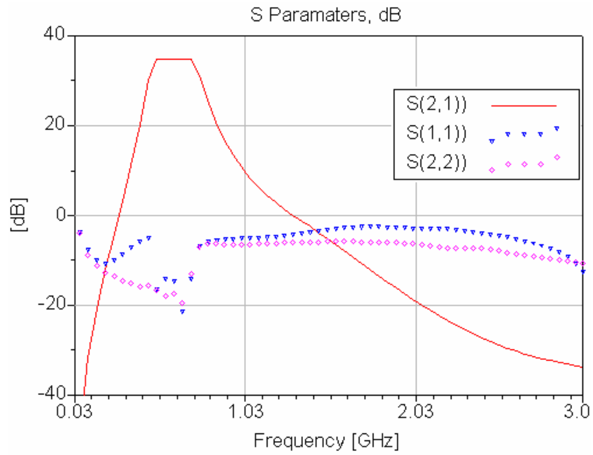

Figure 27 –S parameters simulation results of IF pre amplifier

The layout uses microstrip lines, and the frequencies involved suggested the usual RF PCB design techniques with 0805 or 0603 size SMD components. The PCB board is packaged in an aluminum milled box specially designed for this type of application to serve both as shielding and thermal mass. The layout was designed in ORCAD 9.1 Layout Plus. While designing, the ground connections were placed with the utmost care, putting vias the nearest as possible to the components, the same caution in filtering the supply, introducing two decoupling capacitors for each MMICs supply (100 nF and 100 pF). The free area of the special box to fit the PCB board measures 45 × 16 mm, meaning that the layout was built to satisfy these dimensions. In order to avoid distributed effects due to transmission line usage the electronic components were placed close to each other. This also contributes to reduce the implementation area. The final layout created in ORCAD in its real dimensions is presented in figure 30:

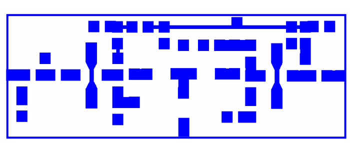

Figure 28 – Final layout of IF pre amplifier

A feed trough capacitor was placed to pass the signal from the power supply to the PCB board. As the term implies, a feed through capacitor has a current-carrying conductor passing through its centre. This co-axial conductor forms one terminal of the capacitor. The other terminal is the metal outer case of the capacitor, which is specifically designed for mounting through the earthed aluminium box. This design feature ensures that any radio frequency currents carried on the central conductor are shunted to earth by the capacitor. After getting the layout complied with all the requirements, it was time to solder all the components. The final circuit for this module is shown in the following picture:

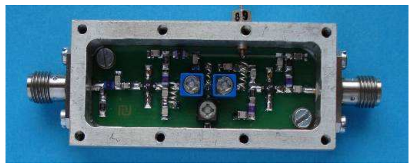

Figure 29 – Photo of IF pre amplifier

It is visible the two ERAs and the components of the filter, the trimmers and the hand made inductors. The bottom of the PCB is obviously the ground plane that is connected to the box. The tests were made using an HP 8753E Network Analyzer and are presented in Figure 32. The power supply was at 8 Volt and the power consumed rounds 400 mW. The ground terminal was attached to the box. This graphic demonstrates the response of the IF Pre amplifier filter after several adjustments of the trimmers, as like the substitution of the inductors, by this way, trial and error, was obtained the desired frequency response and matching results.

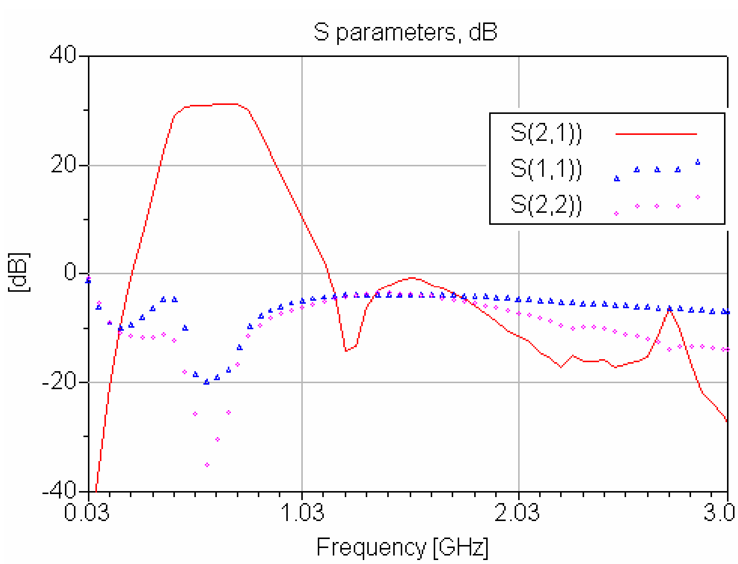

Figure 30 – Measured S parameters of IF pre amplifier

The previous graph displays nice results very close to the simulated. The flat band has variations in the order of 0,1 dB with very good definition of the attenuation and pass frequencies. S11 and S22 show that there are no **Error! Reference source not found.**oscillations (values below -5dB) and that is very well matched in the desired frequencies. The minimum peak near 1 GHz is due to imperfections on the construction of the inductors. Generally this module reports the necessary selectivity of the system (200MHz) at IF (600MHz). The assigned gain in Table I is also accomplished (31 dB). To annotate that because of a mistake in calculating the assumed attenuations for each module in the receiver the first tested module presented a gain of 46 dB. When the error was checked the process of simulating, soldering the components and testing was repeated. The MMICs were two ERA3 instead the present configuration, the filter elements and the slope compensation network had also different values. Briefly, this module was implemented, and tested with pretended values.

2.2.2. IF Amplifier

This stage is the main IF amplifier and will provide most of the gain required. The same procedures while developing the previous module are also applied in this circuit. To keep the good selectivity and sensitivity (2) of the receiver. This module provides very small variations of gain with frequency and very low NF. The same care was taken in avoiding the changes of gain with temperature. The schematic is presented in Figure 33:

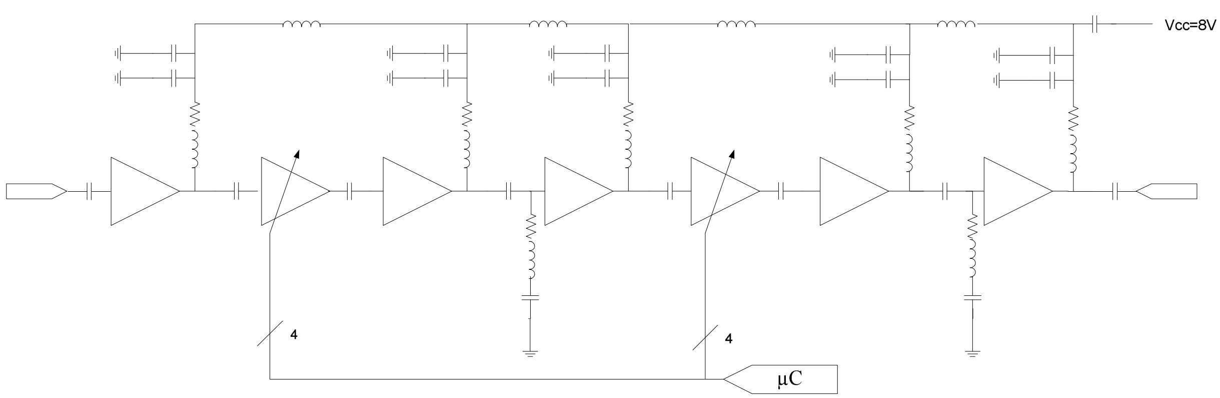

Figure 31 – IF Amplifier Schematic

It is composed of five MMIC stages and two digitally controlled attenuators. Considering the values expressed in Table I it would be required about 56dB of gain, however, by design, it should operate near the middle of gain control settings which is about 15dB below the maximum gain (for a total gain control of 30dB). In this way a total gain of 71dB would be necessary. Controlling the amount of signal at the input of the ADCs is easily accomplished by the MMIC digital RF attenuators DAT-15R5-PP Digital step attenuators from Minicircuits. Each of these has an attenuation of 15,5dB each and is controlled by 5 bits with a 0,5dB step. For Frequency Modulation it is convenient to use this control, to achieve the highest level at the ADCs input. The variation of gain with frequency and temperature was also taken into account, and in this case it would benefit from a very flat band from 100MHz to 1GHz allowing us to have a perfect flatness between 500MHz and 700MHz. The MMIC's chosen were one ERA3 (21dB) and four ERA2 (15,5dB). The expected decrease in gain of the MMIC amplifiers at higher frequencies was again corrected by inserting slope compensation networks, like was done for the IF preamplifier. This time were inserted two RLC networks for the entire module. The attenuator was modeled using simple resistive circuits, which proved to be accurate enough for our needs. Simulation results can be seen in Figure 34, after obtaining S parameters for the ERA3 and ERA2 from the manufacturer.

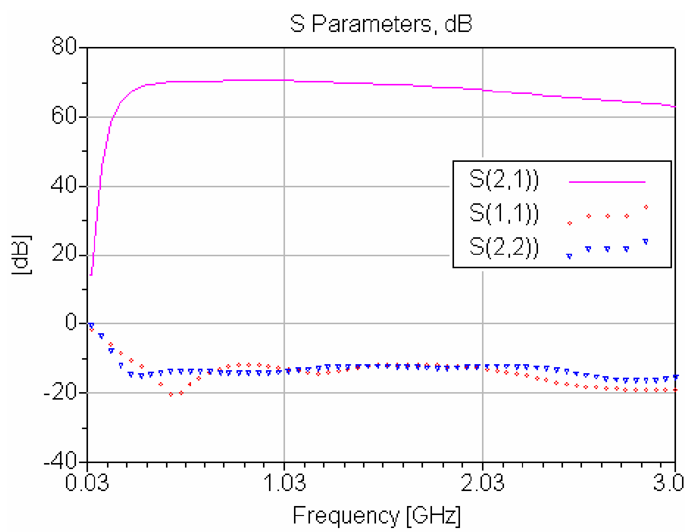

Figure 32 –S parameters simulation results of IF amplifier

Once again this graphic was reached after adjustments of same components, mainly the two slope compensation networks, which load and unload heavily at lower and higher frequencies, respectively. The matched values are has expected, lower than -10 dB extended to a very wide band. The layout for this circuit follows the same lines as for the IF preamplifier.

Microstrip lines were used, and the frequencies involved suggested the usual RF PCB design techniques with 0805 or 0603 size SMD components. The same type of box is used with different dimensions 90 × 16 mm. A Voltage regulator from Torex (XC6202P302MR) was inserted to supply the attenuators (3V). The Bias resistances of the MMICs are like suggested by manufacturer.

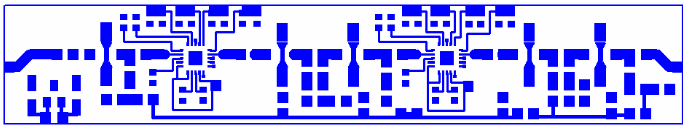

Figure 33 – Final layout of IF amplifier

It is visible the five MMICs and the two digital control attenuators, with its digital inputs, wired wrap was used to connect the inputs. The 3V regulator is on the low left corner, wired wrap was also used to feed the attenuator. Packaging was also approximately the same. The final PCB board already in the special box can be seen in figure 36:

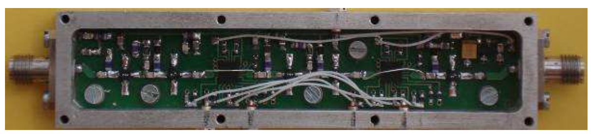

Figure 34 – Photo of IF Amplifier circuit

The test results, obtained with an HP 8753E Network Analyzer are shown on Figure 37.

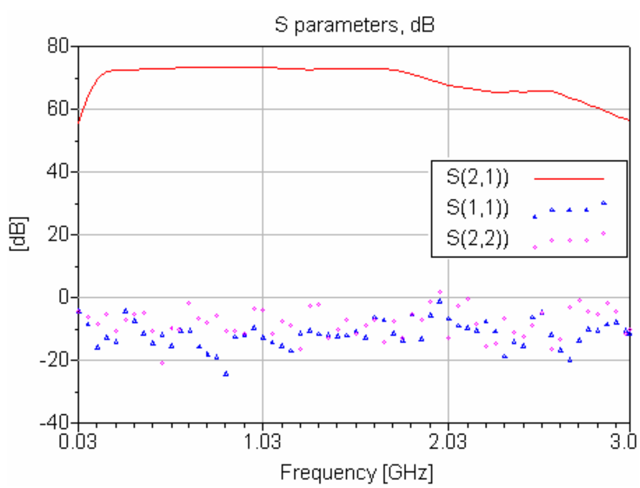

Figure 35 – Measured S parameters of IF amplifier

The tests suggested the need of very good protection against oscillation, in the beginning while submitting this circuit to the Network Analyzer the result was disastrous. Basically the

Analyzer did not retrieve any information. The suspicion was that the circuit was oscillating. A deep investigation of the cause proved that the cover of the circuit served as an antenna and feedback the output power to the input giving rise to an external feedback. This error was checked using an Oscilloscope at the output of all ERAs, and detected that the last two ERAs were oscillating. With a Spectrum Analyzer was determined the Oscillating Frequency at 8GHz. It was also noted that without the cover the circuit did not oscillate, meaning that the cover served as a reflector to the input. To start solving the problem, first was starting to find out the cause that was easily discovered, the cover was not ground shielded, the material of the cover was different from the aluminum box and was not connected to ground. To improve the best isolation between components was decided to insert a cage in each ERA and Attenuator to guarantee the best protection. The budget gain in table I was wrong and like in the IF Pre Amplifier resulted in a maximum gain of 81dB for this circuit. The high gain previewed some problems in developing such a circuit without oscillations, but decoupling the supply with three capacitors (100pF, 4,7nF and 100nF) and rectifying the cover gave rise to a flat gain. The previous description represents some of the problems involved while implementing the circuit, no figures are provided since it were bad results and at the time did not seemed to be important. The first circuit had three ERA3 and two ERA2, to reduce the gain, two ERA3 were replaced by another two ERA2. To obtain a flat gain the Slope Compensation Network was again adjusted by trial and error. A new soldering process to substitute the MMICs and R, L and C was once again done.

The level of the desired signal (~90dBm) to amplify could indicate that noise was not distinguished from signal but this circuit will be placed in a RF Box that controls the temperature in 1ºC. The overall NF for this circuit is basically the NF of the first MMIC, which is typically, for an ERA3, 3,5dB. The NF was computed following (4). The range can be specified in terms of input power or output power. It was also calculated by testing the circuit the compression Point (P1dB). The P1dB was determined testing the circuit for several input powers and for all the attenuation values. The same attenuation was applied in parallel to both the attenuators, the graph below shows the P1dB without attenuation:

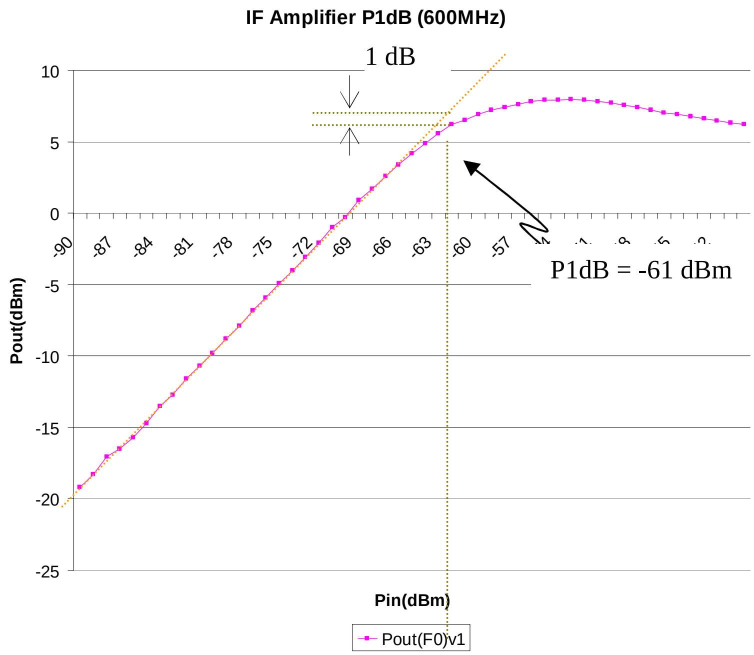

Figure 36 – P1dB point measured of IF amplifier

The P1dB in this case is in terms of input power and is approximately -73 dBm. The same procedure to determine the P1dB for all the attenuations was executed and the values obtained are present in Table 3:

Attenuation (dB)	P1dB (dBm)	Gain
2	-81,6	79,6
4	-79,3	77,5
6	-77,1	75,5
8	-75,1	73,5
10	-73	71,5
12	-71	69,5
14	-68,8	67,5
16	-67	65,8
18	-65,2	63,8
20	-63,1	61,7
22	-60,8	59,6
24	-59,2	57,7
26	-57,3	55,7
28	-55,2	53,7
30	-53,2	51,7

Table 3 – Gain values for several input powers

The previous table describes the P1dB for all the attenuation levels introduced by the two attenuators. It was done with a Network Analyzer pointing out the output power varying the input power at 600MHz. In Table I the gain needed is 56 dB and the input signal rounds -60 dBm, so the ideal operation is with 24 dB of attenuation. The maximum power level for which intermodulation distortion becomes unacceptable is when an input signal consisting of many frequencies result in intermodulation products that may cause distortion of the output signal, this effect is called Third-Order Intermodulation Distortion (IP3). The range can be specified in terms of input power or output power, normally for mixer is referenced to input and for amplifiers to output. For safety it was found the IP3 for this module.

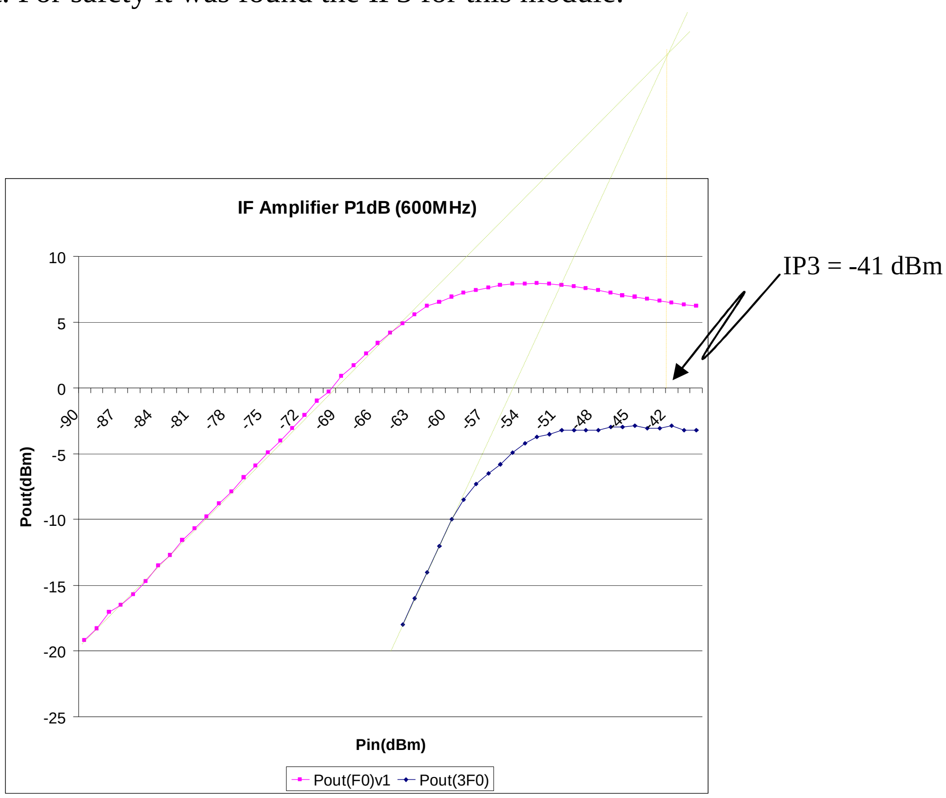

Figure 37 – P1dB and IP3 points measured of IF amplifier

Both responses exhibit compression at high input powers, the plot of the third intermodulation distortion increases quickly than the linear. This module concludes the IF gain of the receiver, it was verified that the saturation point is above the input level signal expected at the input, also the spurious responses are minimal and do not affect the operational region of the IF Amplifier. This module outputs the signal to a Phase and Quadrature Modulator that converts to Zero-IF.

3. Phase and Quadrature Modulation Converter

The Phase and Quadrature modulator is one of the key components, which has significant effects on the quality of modulated signals. This pioneer design exhibits Phase (I) and Quadrature (Q) Modulation allowing the reduction by half the sample rate in digital domain. The 200 MHz bandwidth is centered in base band, so the range is from -100 MHz to 100 MHz, since it is only accessible the range from 0 to 100 MHz this modulation allows having the entire range dispose. This module feeds the Digital Correlator with the required signal level, which is defined by the maximum signal level admitted by the ADC. It is divided in two equal circuits, each having one output that corresponds to an in-phase, or I signal, and the second corresponds to the Quadrature, or Q signal. These allow the preservation of amplitude and phase modulation. It are required two mixers for each circuit, driven by two 600 MHz Local Oscillator (LO) Signals with 7 dBm differed 90º in phase. The phase difference will be made in the cables from the LO to the Converter. The description of this module is divided in two sub-chapters, one for the Converter and other for the LO.

3.1. Converter

Polarimetry is a recent application of digital radiometry to earth science. Microwave polarimetry by digital correlation is based upon the cross-correlation of the horizontally and vertically polarized field amplitude signals where the third and fourth Stokes parameters, are proportional to the I and Q cross correlation, respectively. The advantage to using a digital correlator in polarimetry is that no interchannel signal mixing occurs once the signals are digitized. Detailed information about this is made in Chapter 4. So it is essential to obtain I and Q to determinate the Stokes parameters and this type of converter suites very well to its determination. The signal from each arm of the IF needs to be converted down to a band of frequencies that allows the signals to be processed in the digital domain. Since it is needed to preserve 200MHz bandwidth there are two options: i) converting down to 0 to 200MHz and acquire at 400Ms/s (or slightly higher , in order to account for filter roll off and aliasing issues); ii) converting to zero IF, e. g., to -100MHz to +100MHz acquiring at 200Ms/s (or slightly higher, for the same reasons as above) but in order to preserve the total 200MHz bandwidth a complex signal conversion is required with base-band I and Q signals produced for each arm of the receiver (RHCP – Right Hand Circular Polarization – and LHCP – Right Hand Circular Polarization). By using the second option, the complex conversion to a zero-IF would end up with 4 channels to be digitized but at half the sampling rate that would be required otherwise. This is extremely convenient in order to relax as much as possible the requirements for both the ADC and FPGA in the Digital Correlator. The signals from the two arms of the radiometer are split into its I and Q components. The calculation of the Stokes parameters will become straightforward as this corresponds to the usual rectangular complex number representation. The converter produces the I and Q outputs by multiplying the IF signal with two versions of the local oscillator with a phase difference of 90º. This operation will be applied to both RHCP and LHCP arms, so the implementation uses two identical circuits.

The 600MHz IF signal is separated into two channels, using a power splitter ADE-2-9 from Minicircuits to feed a pair of ADEX-10 mixers. The local oscillator (LO) already outputs the correct driving level for the mixers, 7dBm, and the correct phase relation, 0 and 90º of phase difference made in the cables. Once the I and Q signals are converted it is convenient to eliminate possible spurious caused by the mixer down conversion, for that the I and Q signals obtained will be low pass filtered to remove unwanted frequencies, then amplified and finally low-pass filtering is combined to reject the images and suppress out of band spectrum prior to digitization. A diagram of the converter is shown on Figure 40.

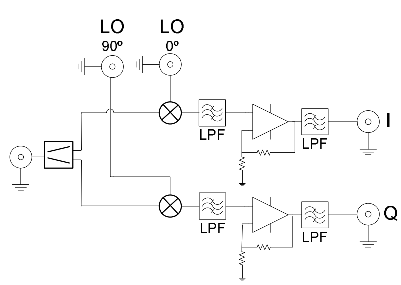

Figure 38 – Zero-IF converter block diagram

The filters that follow the mixers are 7th order Butterwoth low-pass filter with 100MHz cut-off frequency. The output filters are 3rd order Butterworth low-pass with 120MHz cut-off frequency. This last filtering prior to the digital acquisition of signals will eliminate any wideband noise and spurious signals that might be presented at this point ensuring that only the signals of interest are presented to the ADCs.

Both filters were implemented using lumped L and C elements. The filter was simulated in a design guide tool from ADS (Filter Design Guide) for lumped components, it was only needed the band /stop band frequencies and the band/stop band attenuation values. A final adjustment on the filter elements created by the ADS tool was executed to achieve the wanted filtering. The S21 parameters describe the transfer function of the filter, like is shown in figure 41:

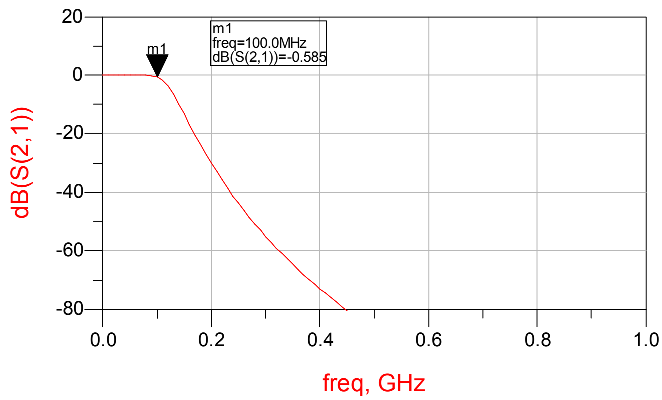

Figure 39 – S21 parameters simulation result of filter 1 in converter

To avoid the excess use of inductors, the second filter is simpler than the first, instead of π type, is a T type. It also has a 100MHz cutoff frequency, but a low order (N=3). Its S21 parameters are in figure 42.

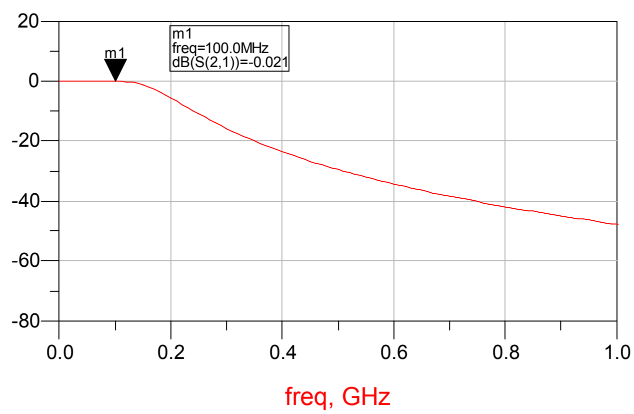

Figure 40 – S21 parameters simulation result of filter 2 in converter

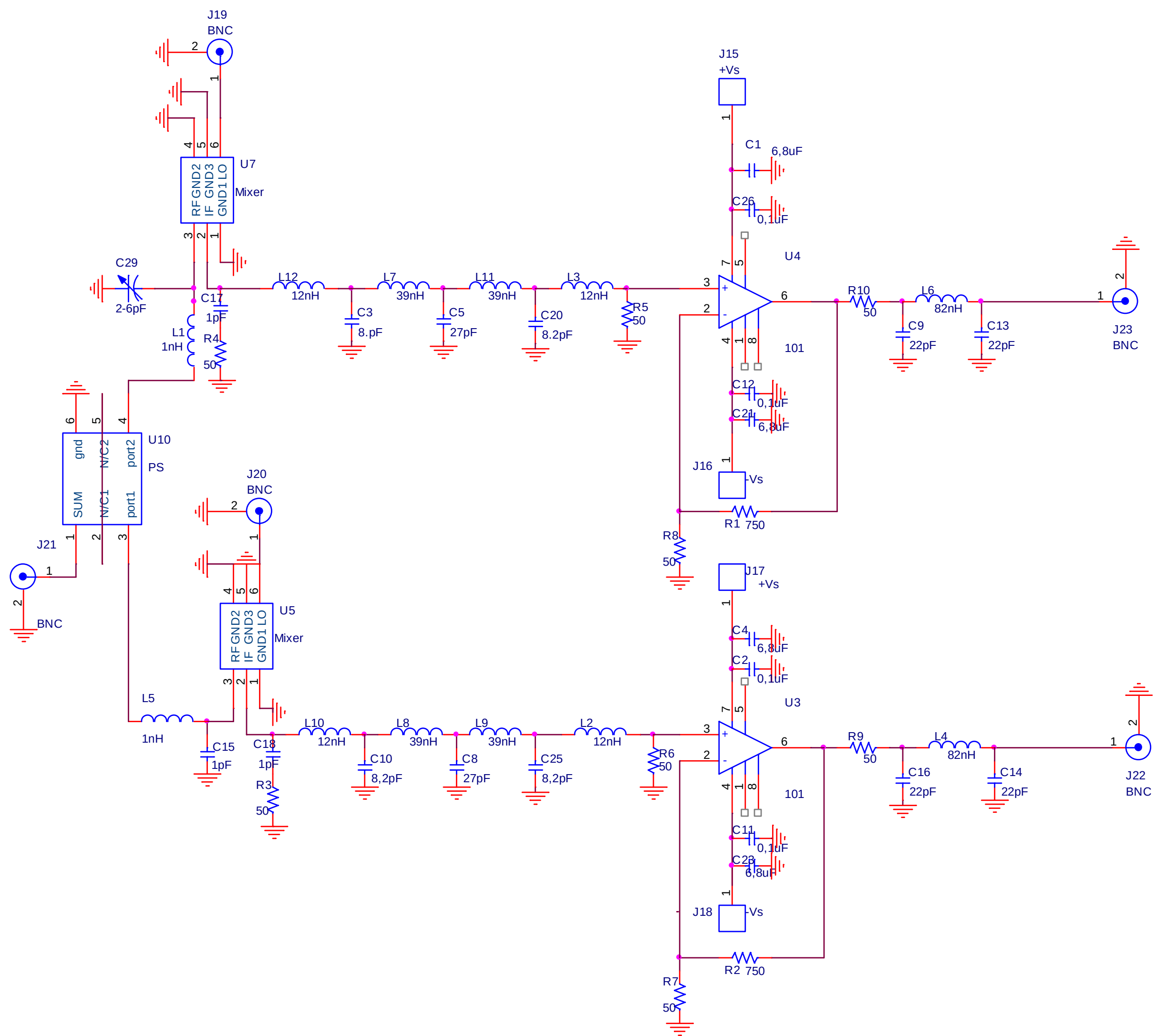

Figure 41 – Schematic from second converter

The amplifier uses the OPA657, an OPAMP (operational amplifier from Burr-Brown) in a non-inverting configuration. The market options for 1 GHz Gain Bandwidth Product (GBWP) OPAMP imposed this choice. This way is secured a 20 dB gain at 100 MHz. But OPA657 has a 1,6 GHz GBWP, and by datasheet specifications for a 1Vpp output voltage at 100MHz the gain is near 21 dB which is settles perfect to the needs. The layout follows the usual techniques for RF instrumentation using both PCB layout and external metallic shielding. Further care was necessary to externally ensure phase balance both in the RF and LO signal path, in order to guarantee perfect orthogonal output signals in all circumstances. Fine phase trimming was provided with a variable capacitor to allow a precise calibration.

The circuit employs microstrip layout design and is constructed on a FR4 epoxy substrate. The final preview of the PCB design is presented in the figure below:

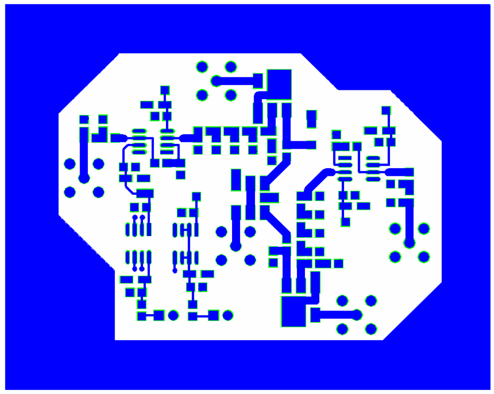

Figure 42 – Final layout of I and Q modulator

In order to guarantee equal output signals the lines needed to be very carefully designed, to assure this lines having the same length were implemented and with the help of the phase trimming. The area that involves the circuit is a ground shield to isolation. The manufactured PCB having these characteristics is in figure 45, the bottom plane is ground:

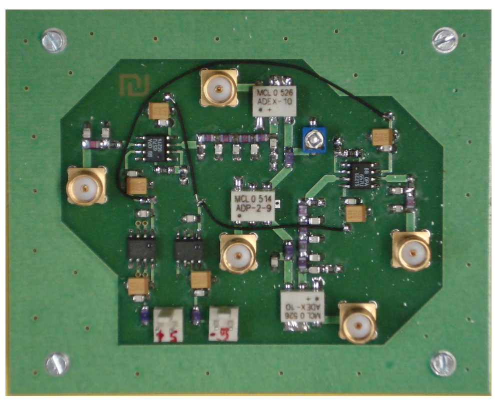

Figure 43 – Photo of I and Q modulation converter

For safety the circuit area will be covered by a metal plaque connected to the ground shield, all the connectors will stay on the bottom plane. This way the converter is kept clear from any strong signal that could enter the RF port in the mixer. The isolation between ports was important in the selection of the mixer. Isolation is 120 dB between ports. For the tests were used an RF signal of 600 to 700MHz with -7dBm and an LO signal at 600MHz with

7dBm. The output signal, varying from DC to 100MHz was measured with a spectrum analyzer and presented in Figure 446.

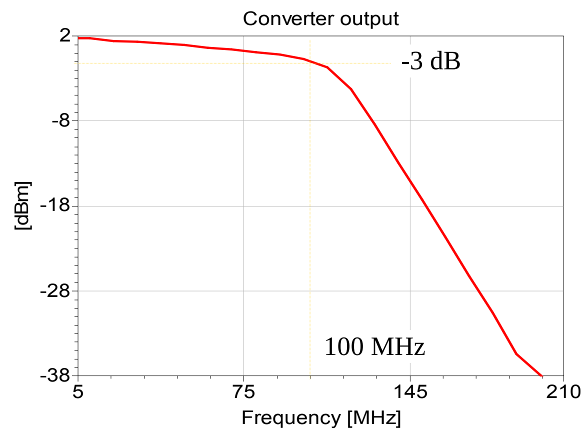

Figure 44 – Converter measurement results as measured on a HP8563A spectrum analyzer

The I and Q signals from both arms of the receiver will be outputted to the digital correlator like a shape identical to the described in the graphic of Figure 45. The necessary 100 MHz bandwidth is defined and presenting the gain determined in Table I to feed the four ADCs. The down conversion is provided by an LO implemented for this work which is described in the next section.

3.2. Local oscillator

This circuit is responsible to drive the mixer in the converter with a strong signal that is generated by a voltage tunable oscillator and synthesized by a frequency synthesizer. Its function is to drive the components of the mixer into a nonlinear regime for frequency mixing. Phase noise is an important specification of oscillator, since any phase fluctuation is overlapped on the mixer output signal. Isolation is once again a feature carefully applied.

This circuit provides the LO signal to the converter. Since it is wanted to do a zero-IF conversion, the required oscillator frequency is 600MHz, the exact central frequency of the IF pass-band. It was decided to use the encapsulated oscillator ROS615 from Minicircuits which is a VCO (voltage controlled oscillator) that tunes from 580 to 615 MHz. It is operated at the fixed frequency of 600.0MHz. There is both the possibility to tune the frequency using a potentiometer, varying the voltage from 0V to 5V or it can be used a PLL synthesizer chip

LMX2326 from National Semiconductors. As for the moment no frequency stability better than 1MHz is required, so the LO operates with analog frequency control.

The oscillator signal needs to be split into four equal signals. For simplicity it was accomplished this with simple resistive dividers in T attenuator configurations allowing for normalization of the gain and attenuation required, thus providing four identical signals close to 7dBm. Since the signal from the VCO was considerably low (about 0dBm) it was needed to amplify it before splitting the signal. In order to have isolation between the output ports of this unit it is desirable to have one amplifier per output. Taking these requirements into account it arrived the design presented on the diagram of figure 47.

It was used an ERA1 (from Minicircuits) MMIC amplifier to raise the power of the VCO to about 10dBm. The resistive divider then lowers each output power to about -4dBm. For this reason another ERA1 amplifier is needed to raise output power again to 7dBm, the power required to drive the converter block.

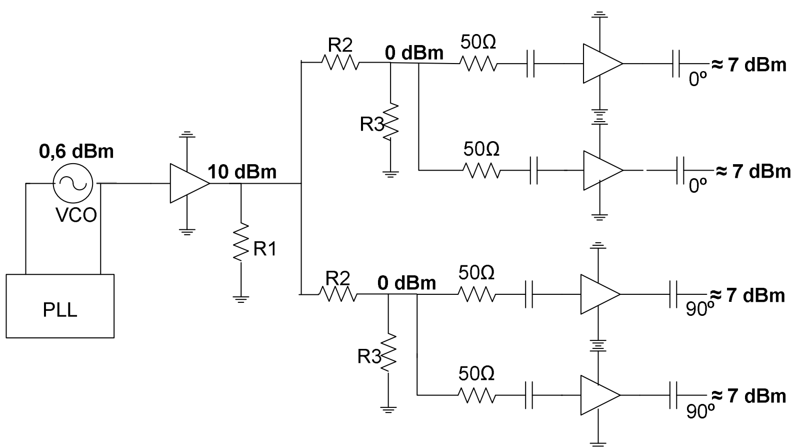

Figure 45 – Local oscillator module block diagram.

The same layout considerations and RF common practices apply for the local oscillator module. The same protection in the converter is again applied here, because any strong signals in the receiver chain could enter this unit damaging the output signals. The paths have the same length to assure four equal output signals, to realize the 90º phase difference the cables that connect to the converter have different lengths.

This block was also implemented on FR4 substrate using microstrip design and SMD components. A preview of layout and the manufactured PCB board are presented in Figure 468.

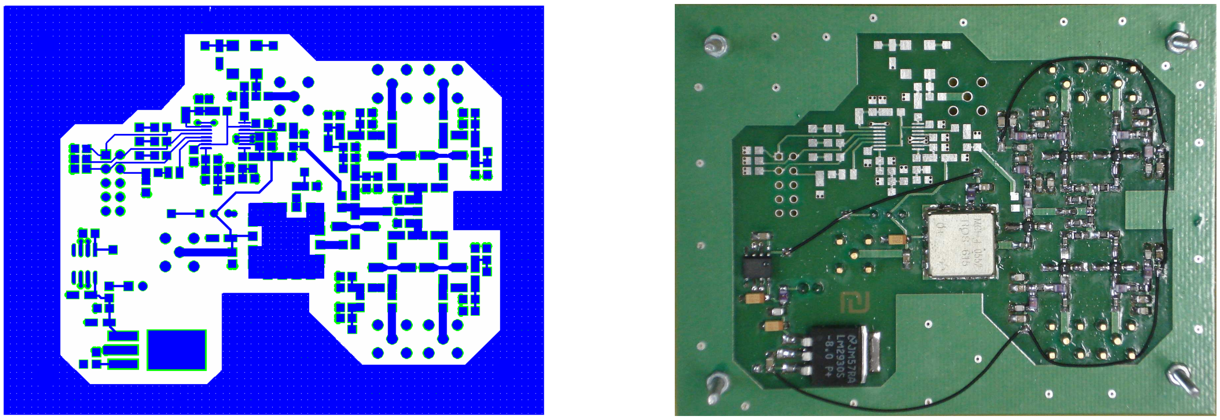

Figure 46 – Layout and photo of final LO circuit

The same layout considerations and RF common practices apply for the local oscillator module. This block was also implemented on FR4 substrate using microstrip design and SMD components. All the connectors are placed in the ground plane at the bottom layer. Again the supply connections are made by wires. The schematic of this module is in figure 49:

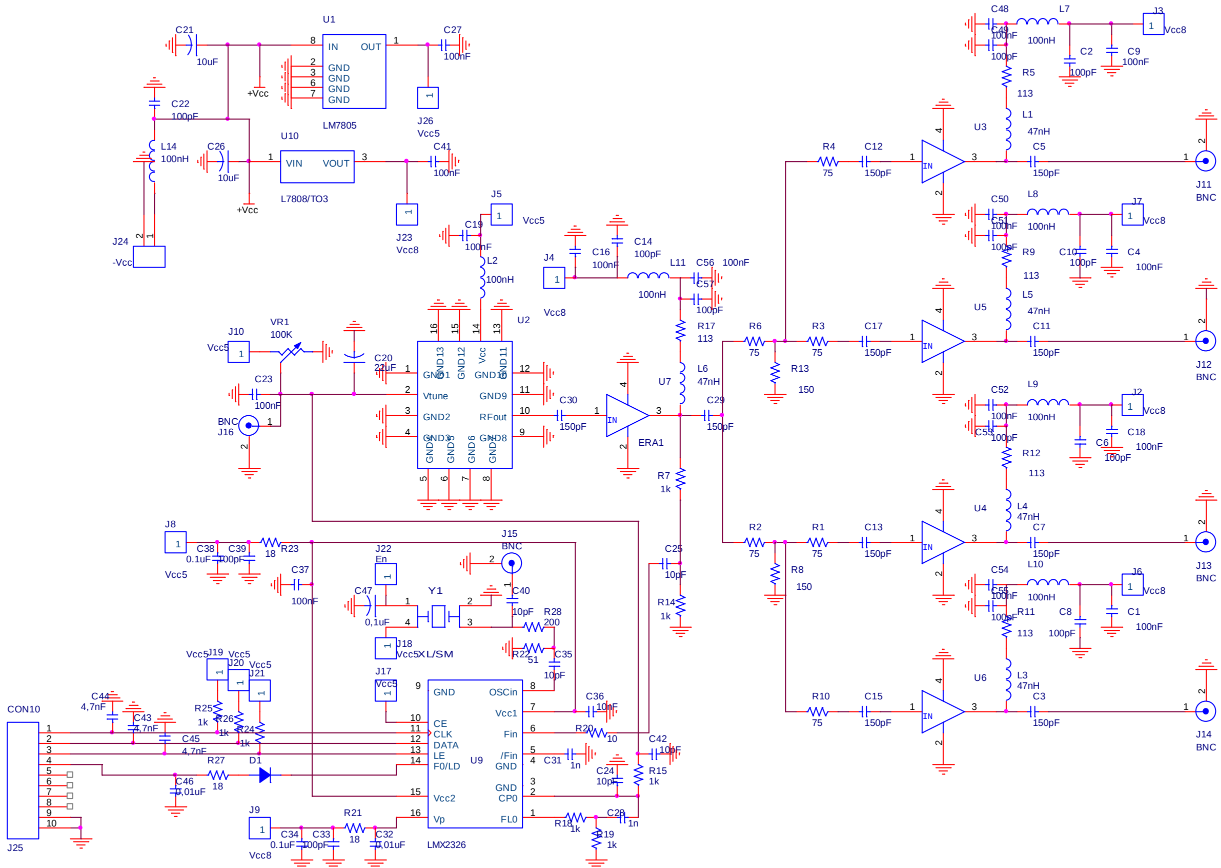

Figure 47 – Schematic of LO

4. Full Digital Correlator

This module performs the digitization of the signal and also the autocorrelation/integration of the digital data. The bandwidth of the four incident signals are now 100MHz at zero IF. To satisfy the Nyquist principle the ADCs need to sample twice the speed of the signal bandwidth, that is 200MHz. To fulfill these requirements it is convenient to use flash ADC with the capability of interleaving data. Interleaving is the process of break in two (or more ways) a result at the same processing speed, reducing the sample rate by half (or less). The main operations of this module are described in a diagram in figure 50.

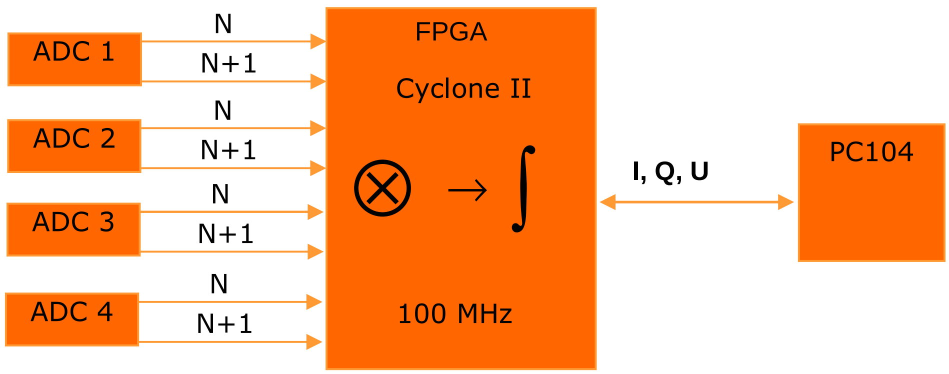

Figure 48 – Block diagram of Full Digital Correlator

The correlations are made in an FPGA and are very useful to calculate the Stokes Parameters, also in the FPGA. This circuit is composed by four ADCs and one Altera Cyclone II FPGA. The chosen ADCs accept a 0,5 V peak-to-peak signal and the amplitude signals stay between half of the range and the maximum allowed of the ADCs input signal through the help of digital control attenuators. Since their best performance occurs for differential signals, a differential amplifier was inserted to transform the single ended signal in a differential one. The digital signals have already been interleaved, internally to the ADCs, so each ADC will present two times eight bits data outputs at each cycle corresponding to sample n-1 and sample n. allowing a slower clock but more massive parallelization. The speed of the FPGA is then half of the sample rate that is 100MHz while the ADCs run at 200MHz.

The polarization of the radiation gathered by the radiometer is best described calculating its Stokes parameters.

$$\text{Stokes U} \rightarrow RL = \Re e\left(E_{rcp} E_{lcp}^{*}\right) \tag{13}$$

$$\text{Stokes Q} \rightarrow LR = \Re e\left(E_{rcp,-\frac{\pi}{2}} E_{lcp}^{*}\right) \tag{14}$$

$$\text{Stokes I} \rightarrow RR + LL = \left\langle E_{rcp} E_{rcp}^{*} \right\rangle + \left\langle E_{lcp} E_{lcp}^{*} \right\rangle \tag{15}$$

To obtain these parameters the digital data (samples) from ADC are correlated and integrated, in parallel, inside the FPGA. After processing and integrating several samples the data rate is slow enough to be transferred to a computer. This computer, a PC104 module (AMD LX800 500MHz from Kontron), interfaces the FPGA by the ISA bus. This PC is responsible to format the data store it locally and make it available on the network via Ethernet IP using SFTP (secure file transfer protocol). All the configuration of the FPGA is being developed using VHDL language on Quartus II v6.1 web edition software. The four layer PCB layout of FPGA obeys no specific standard form. High speed digital design considerations were applied during this design. The layout is in the following figure:

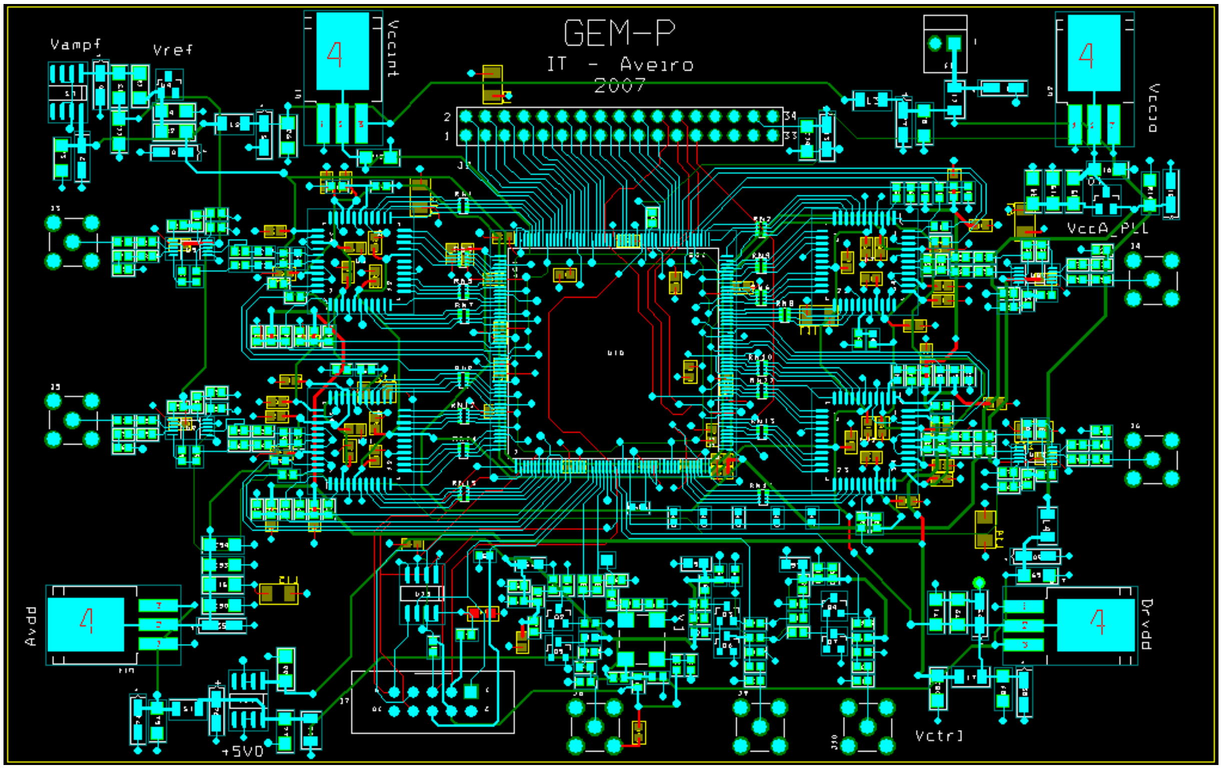

Figure 49 – Final Layout preview of Digital Correlator

5. Conclusions

To accomplish the polarimetric measurements at 5GHz for the Galactic Emission Mapping in the North Hemisphere was developed a novel instrument with all polarimetric measurements being performed in the digital domain. In order to achieve this objective, a heterodyne radiometer / polarimeter is being developed and a new intermediate frequency chain with high performance had to be designed along with a new digital correlator implemented in an FPGA developed in VHDL language.

This thesis addressed common design problems found in the project of radio astronomy electronic receiving equipment. Amongst these problems the design of IF Amplifier and I-Q modulation converter are considered of paramount importance, since they impair the overall system performance. The IF amplifier was carefully developed in order to achieve the highest gain on the receiver chain 71 dB, for that were used five MMICs together with two attenuators to control de signal level at the ADC input. Due to normal decrease of gain with frequency were inserted two slope compensation networks to obtain a flat gain during a large bandwidth. Some protection considerations were taken into account in order to avoid any stage oscillation. While testing some oscillations occurred but were solved grounding the cover of the alumina box. The IF amplifier is ready to be applied to the receiver. In the I-Q modulation converter is suitable of the calculation of the Stokes Parameters, since its phase and Quadrature components are the elements of the Stokes equation parameters. Each polarization is treated equally and divided in two arms, each arm have all the same line length. The mixer has low conversion loss and high isolation between ports, again to protect from other signals. A Diplexer filter is inserted to match the highest frequencies and filter the desired 100 MHz bandwidth. Finally to provide the analog to digital conversion with the needed value there is signal amplification on a non inverting configuration

Filtering at RF is very difficult with active devices so for this case were applied microwave concepts to develop a fine selectivity at 5 GHz. Coupled showed to be the best choice, like demonstrated in the results, it also served to know the tool Momentum from ADS, that perform high performance 2,5D Electromagnetic simulation.

With no less importance there are the other stages implementation, namely, the IF pre amplifier, that has the special characteristic of selectivity of the receiver along with a pre amplification to provide the IF amplifier. The same considerations were applied like in the IF amplifier.

The LO provides the Converter with a strong signal to multiply by the IF signal for zero-IF conversion. The frequency synthesize is made by a simple potentiometer tuned at 3,25 V for 600 MHz or for a fine tuning by a PLL not yet implemented. To feed the converter with 7 dBm, the signal is then divided in four outputs and amplified by one MMIC. The phase difference is applied in the length cables to the converter.

Analog to digital conversion, correlation, integration ad Stokes parameters calculations are made in the digital correlator. This stage performs all the operations at 100MHz for the FPGA and 200 MHz for the ADCs. The hardware design secures the maintenance of properties of the four signals, like is verified in the layout, it also outputs the calculated data to an outside computer.

To conclude, the entire IF system and the digital correlator was implemented using standard SMD components and classical approaches to high frequencies and microstrip design, but using commercial RF MMIC devices. Therefore, a high performance IF strip attended by a digital correlator, suitable for a radio-astronomy application, and in particular for the GEM (at Portugal) experiment, was designed, constructed and tested successfully.

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